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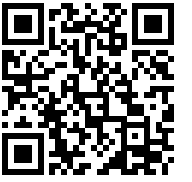

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VOLUME 12

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RECEIVING MEASUREMENTS AND ATMOSPHERIC DISTURBANCES AT THE RADIO PHYSICAL LABORATORY, BUREAU OF STANDARDS, WASHINGTON, JULY AND AUGUST, 1923*

By

L. W. Austin

(Chief of the Radio Physical Laboratory, Bureau of Standards, Washington, D. C.)

(Communication from the International Union of Scientific Radio Telegraphy)

The observations for July and August afford some opportunities for a comparison of this summer's result with the corresponding values of last year.

Lafayette's change of frequency from 12.8 kilocycles per second (kc.) to 16.2 kc., mentioned in the May-June report, has been accompanied by an increase in signal strength in Washington of about 1.6 times. As far as is known, there have been no changes other than the change in frequency to explain the increase.

This, together with some observations on shorter wave stations, may indicate a distinct optimum frequency in the neighborhood of 16 kc. for northern trans-Atlantic communication, but of course it is still too early to draw definite conclusions on this point.

The signal strength of Nauen, both in the morning and during the afternoon fading, is practically the same as in the corresponding months last year. The atmospheric disturbances are considerably lighter than in 1922.

*Received by the Editor, November 3, 1923

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FIELD INTENSITY OF NAUEN AND OF DISTURBANCES (f=23.4 kc. λ=12,800 m.) IN JULY, 1923, IN MICRO-VOLTS PER METER

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
2	34.7	30	6.5	150
3	47.0	30	2.0	210
5	30.0	40	*	135
6	23.4	30	*	125
7	16.3	42	....	—
9	30.0	25	*	160
10	19.0	35	*	180
11	23.4	30	*	300
12	34.2	20	2.0	200
13	22.0	22	*	180
14	17.0	30	**	....
16	13.7	100	*	380
17	55.7	48	16.1	240
18	42.0	60	18.0	60
19	36.4	48	18.4	150
20	47.0	65	12.2	100
21	**	60	....	....
23	55.7	48	2.0	150
24	42.7	48	*	240
25	42.0	65	*	380
26	72.5	30	27.0	100
27	....	....	....	....
28	....	....	....	....
30	**	28	*	360
31	**	15	6.3	130
Average	35.2	41.2	5.5	196.5

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FIELD INTENSITY OF LAFAYETTE AND OF DISTURBANCES ($f = 15.9$ *kc.* $\lambda = 18{,}900$ m.) IN JULY, 1923, IN MICRO-VOLTS PER METER

Date	9 A. M.		3 P. M	
	Signal	Disturbance	Signal	Disturbance
2	**	40	66.3	200
3	126.0	35	40.4	260
5	120.0	50	24.5	135
6	66.3	35	2.0	180
7	84.0	60	....	....
9	108.0	35	15.0	170
10	120.0	55	*	210
11	102.0	48	42.3	350
12	137.0	30	57.3	280
13	140.0	25	36.2	200
14	120.0	40	....	....
16	77.2	160	2.0	410
17	330.0	60	102.0	260
18	108.0	50	84.0	70
19	160.0	65	114.0	150
20	160.0	80	84.0	130
21	90.0	75	....	....
23	168.2	80	45.2	175
24	156.2	65	18.0	280
25	150.0	55	2.0	350
26	175.0	80	60.0	150
27	....	....	....	....
28	....	....	....	....
30	150.0	35	2.0	550
31	66.3	20	40.2	150
Average	132.4	55.5	41.9	232

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FIELD INTENSITY OF NAUEN AND OF DISTURBANCES (f=23.4 kc., λ=12,800 m.) IN AUGUST, 1923, IN MICRO-VOLTS PER METER

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
1	26.0	10	10.5	150
2	25.7	65	10.0	200
3	**	20	....	....
4	26.2	60	....	....
6	25.6	35	*	300
7	25.0	55	....	....
11	25.6	50	....	....
13	25.6	25	15.0	200
14	38.2	40	*	110
15	25.6	60	2.0	200
16	34.2	50	*	250
17	26.6	45	*	200
18	25.7	30	....	....
20	47.0	30	9.3	100
21	21.5	20	*	280
22	27.0	30	*	250
23	30.0	20	2.0	80
24	25.7	48	30.0	120
25	25.7	40	....	....
27	30.2	25	2.0	240
28	27.0	60	9.0	250
29	**	50	*	280
30	18.0	65	*	200
31	31.0	50	*	500
Average	27.8	40.9	5.0	217.1

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FIELD INTENSITY OF LAFAYETTE AND OF DISTURBANCES ($f = 15.9$ kc., $\lambda = 18{,}900$ m.) IN AUGUST, 1923, IN MICRO-VOLTS PER METER

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
1	102.0	15	60.2	180
2	84.0	85	90.0	250
3	152.0	25	....	....
4	84.0	65	....	....
6	60.0	50	72.2	320
7	78.2	75	60.0	300
11	102.0	80	....	....
13	78.2	35	54.0	250
14	78.2	35	42.2	110
15	120.0	60	30.4	250
16	**	65	18.0	300
17	84.0	60	2.0	250
18	145.0	45	....	....
20	152.0	35	60.2	110
21	126.0	25	25.4	300
22	84.0	30	30.0	300
23	96.0	25	96.0	100
24	152.0	45	120.0	150
25	126.0	50	....	....
27	120.0	25	48.3	240
29	**	60	30.2	300
30	120.0	80	40.0	250
31	120.0	60	2.0	630
Average	107.7	49.1	48.9	255

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f kilocycles	Signal P. M. / A. M.	Disturbance P. M. / A. M.	A. M. Signal / Disturbance	P. M. Signal / Disturbance
		JULY		
15.9	0.316	4.18	2.38	0.180
23.4	0.156	4.77	0.854	0.028
		AUGUST		
15.9	0.454	5.20	2.19	0.192
23.4	0.180	5.31	0.680	0.023

Radio Physical Laboratory,
Bureau of Standards,
September 25, 1923.

SUMMARY: The signal strengths of the Nauen and Lafayette Stations, and the corresponding strength of the atmospheric disturbances in Washington are given for July and August, 1923.

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THE RADIO EQUIPMENT OF THE STEAM YACHT "ELETTRA"*

By

Eric A. Payne

(Marconi's Wireless Telegraph Company, Ltd.)

The steam yacht "Elettra" was designed by Messrs. Cox and King, of London, and built by Messrs. Ramage and Ferguson, at Leith, Scotland, being completed in May, 1904. She was originally the property of the Archduchess Maria Theresa of Austria, being then called "Rovenska," and was used by the British Admiralty for naval purposes during the war. In the year 1919, she was purchased by Senatore Guglielmo Marconi and renamed "Elettra."

The principal dimensions are as follows: Overall length, 220 feet (72.2 m.), length at water line, 198 feet (64.9 m.), beam, 27½ feet (9.2 m.), and the depth, 16½ feet (5.4 m.). The registered tonnage is 232, gross tonnage 633, and the Thames measurement tonnage 693. The yacht is driven by a triple expansion condensing steam engine. The high, medium and low pressure cylinders have diameters of 16, 26, and 42 inches, respectively, and the stroke is 27 inches. The nominal horsepower is 137, and the indicated horsepower is about 1000.

There are two coal-fired boilers having grate surfaces of 95 square feet area, and a total heating surface of 645 square feet.

The working pressure for the main boilers is 180 pounds per square inch, and for the donkey boiler 120 pounds per square inch. The normal cruising speed is 10 knots, with the engine running at 90 revolutions per minute. The consumption of fuel at this speed is about 12 tons per day, and a sufficient quantity may be carried for about 12 days, normal steaming.

Sails are also used when the wind is favorable, as, apart from the speed gained, they assist in steadying the yacht. Figure 1 is a view of the yacht, and Figure 1A shows the arrangement of the antennas.

The main storage battery for lighting and supplying the radio

*Received by the Editor, August 17, 1923.

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when the dynamo is not running is situated in the propeller shaft tunnel. It consists of 36 cells and has a capacity of 500 ampere-hours. This battery is charged by a dynamo which is driven by a small steam set. A gasoline motor is also provided for use when the yacht is in port and there is no steam available.

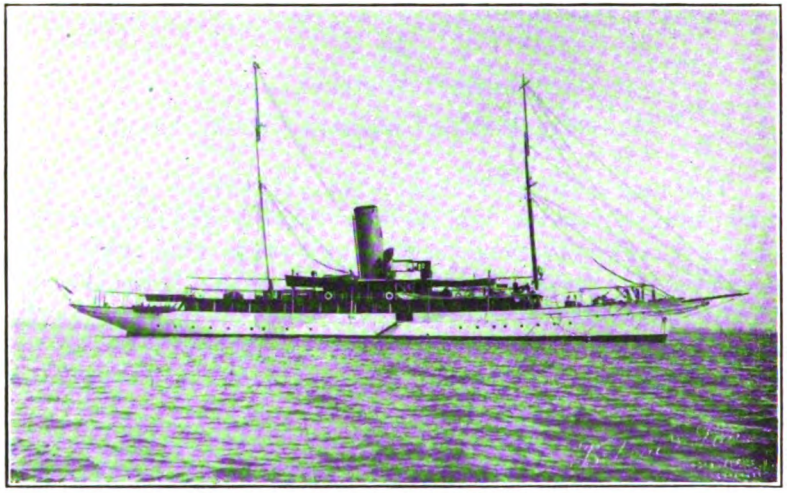

FIGURE 1—The Steam Yacht "Elettra"

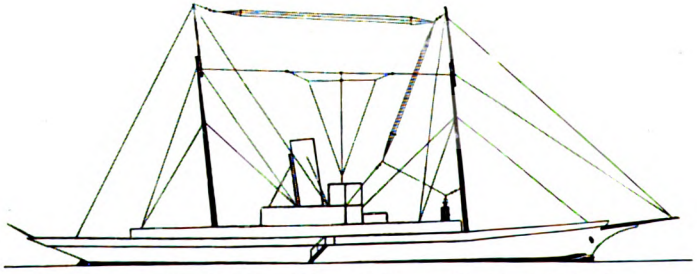

FIGURE 1A—Rigging Plan of Steam Yacht "Elettra"

A submarine signal receiver is fitted with which it is possible to tell from which direction the signal is coming. There are two microphones situated inside the hull, one on either side of the bow. By means of a throw-over switch, either of these may be connected to a battery and telephones located in the chart house. When in the vicinity of a signal station, the navigator listens in the telephones, and the microphone which is more directly facing the signal bell gives the louder response. The ship's log has an electric repeater, whereby the distance travelled thru the water

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is recorded on dials in the engine room and in the chart house. A powerful searchlight is fitted to the foremast, the light being supplied by an electric arc lamp.

The yacht is navigated by a Sperry Gyroscope compass. The master gyro compass is installed below in a fairly central position, and the charging generator and motor alternator for revolving the gyro rotor, are placed in the engine room. The switchboard for controlling the various circuits is fitted near the master compass.

One electrically operated repeater for steering purposes is located on the bridge, and another on the top deck for taking bearings of distant objects. By the use of this compass in conjunction with the radio direction finder, very exact navigation is effected. Figure 2 shows the yacht being steered by Sperry gyro compass.

FIGURE 2—Steering by Sperry Gyro Compass

VALVE TELEGRAPH TRANSMITTER

This transmitter is worked from alternating current which

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is supplied by a motor alternator driven from the yacht's main lighting battery. The supply is 400 volts at a frequency of 300 cycles. The filament current is supplied by small step-down transformers and is regulated by chokes with variable iron cores, operated thru gearing, by handles from the front of the transmitting panel. For the high tension plate voltage, the alternating current supply is stepped up by an oil-cooled transformer to 12500 volts, and then rectified by two 2-electrode valves. This transformer has a centre point tapping, and the valves are arranged for double rectification. A condenser of large capacity is connected across the output, and iron cored chokes are placed in series, and a fairly smooth and steady supply is obtained, of direct current at about 8000 volts. Figure 4A shows the connections of the rectifier circuits.

This direct current, high tension supply is fed to three 3-electrode valves, which have their plates connected in parallel. Each grid is connected to the reaction coil thru a separate condenser and resistance, this reaction coil providing a variable coupling between the plates and grids. The plates are connected to the aerial inductance thru a blocking condenser, composed of four glass tube condensers. In series with the direct current high tension supply to the plates is an air-cored choke and between the supply side of this and the earth connection, another glass tube condenser is connected, for by-passing any high frequency currents which may have got thru the choke.

Signaling is effected by simultaneously breaking the main transformer primary, and the grid circuit, by means of a relay, which is controlled by the hand-telegraph key, and operated by the direct current supply. Figure 4B shows the oscillator circuits.

The handle of the safety gates of the transmitter is mechanically coupled to a switch, so that the power is cut off when these gates are opened. Also, the act of opening these gates operates push rods, which discharge the smoothing condensers, so that the apparatus is safe to handle.

Ammeters are provided for reading the filament and plate currents. The power input to the transmitter is regulated by an iron-cored choke, connected in the primary circuit of the main transformer. To compensate for the drop in voltage caused when the key is pressed, a small iron-cored compensating choke in the primary circuit of the filament transformers is short-circuited by an additional contact on the signaling relay. By means of a variable iron core, the value of this choke may be

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adjusted so that the filament brilliancy is maintained constant during telegraphic transmission.

The amount of coupling between the grid and plate circuits, and the position of the plate connection to the aerial inductance, are adjusted to give maximum aerial current, at the same time, keeping the plate current as small as possible.

Any wave length from 200 to 3000 meters may be obtained by means of a variable inductance and series condenser; a rough adjustment being made by a tapping and the final adjustment by a variometer.

Telephony

The same transmitter may be used for telephonic communication when required. Modulation is effected by two valves, which have their plates connected to the high tension, direct current supply. The grids of these valves are coupled thru an iron-cored choke and battery to the filaments, and also thru a coupling condenser, to the plate of an amplifying valve.

The plate current for this amplifying valve is also supplied from the same high tension supply, thru a large resistance, and its filament current is supplied from an accumulator, so as to ensure an absolutely steady emission.

A microphone and battery are connected to the grid of this amplifier valve thru a small iron-cored transformer.

Ammeters are connected in the plate circuits of both modulating and amplifier valves, so as to indicate the amount of modulation.

The telephone transmitter is operated by a remotely controlled relay, which changes the aerial from the receiver, lights up the modulating valves, and starts the main transmitter oscillating.

In order to carry on conversation with another station, it is necessary to operate only one small switch, which controls this relay. Figure 3 shows the front of the transmitter, Figure 4 the rear view with the control relays, Figure 4C the modulator, and Figure 5 the complete layout on the yacht.

Main Spark Transmitter

Power for the main spark transmitter is supplied by a motor alternator, driven from the direct current supply from the yacht's battery. Alternating current of 1½ kilowatts is supplied at a voltage of 200, and a frequency of 500 cycles.

This voltage is stepped up by an oil-cooled transformer and

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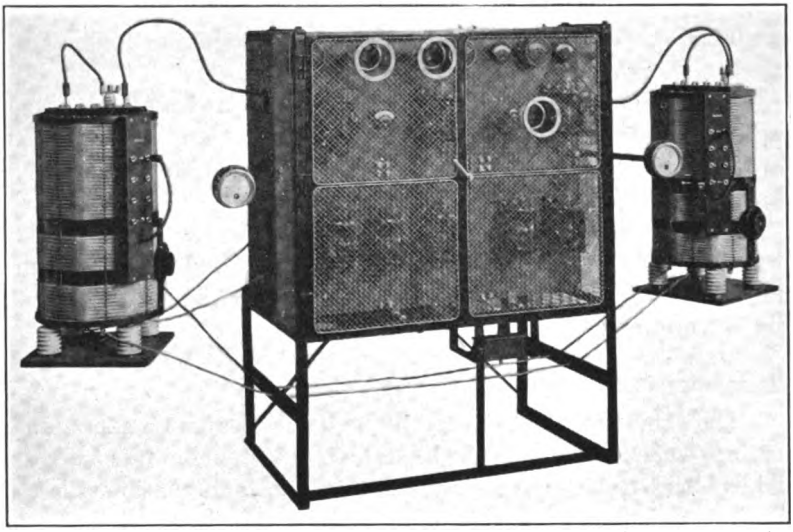

FIGURE 3—3 K.W. Valve Telegraph and Telephone Transmitter, Arranged for Inductive Coupling to Aerial

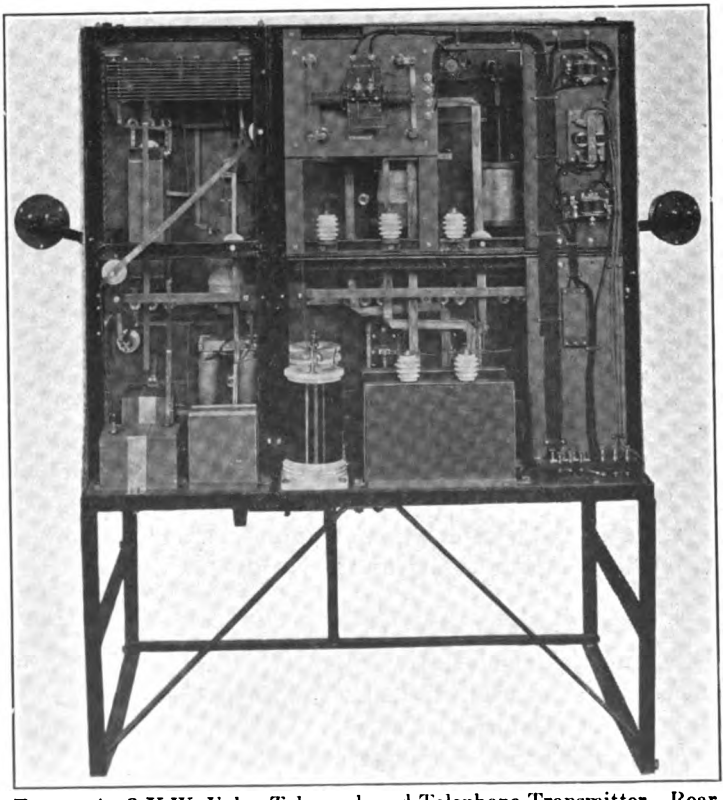

FIGURE 4—3 K.W. Valve Telegraph and Telephone Transmitter—Rear View Showing Control Relays

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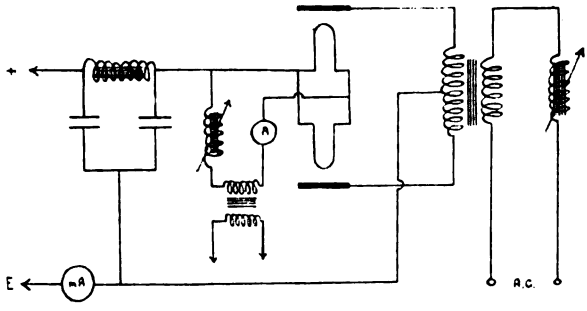

FIGURE 4A—Rectifier

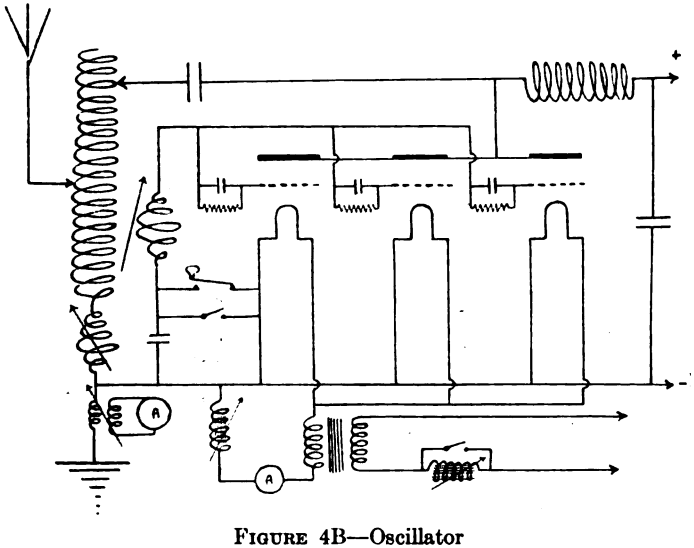

FIGURE 4B—Oscillator

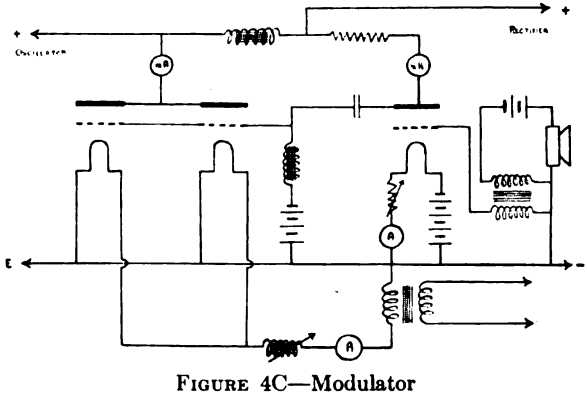

FIGURE 4C—Modulator

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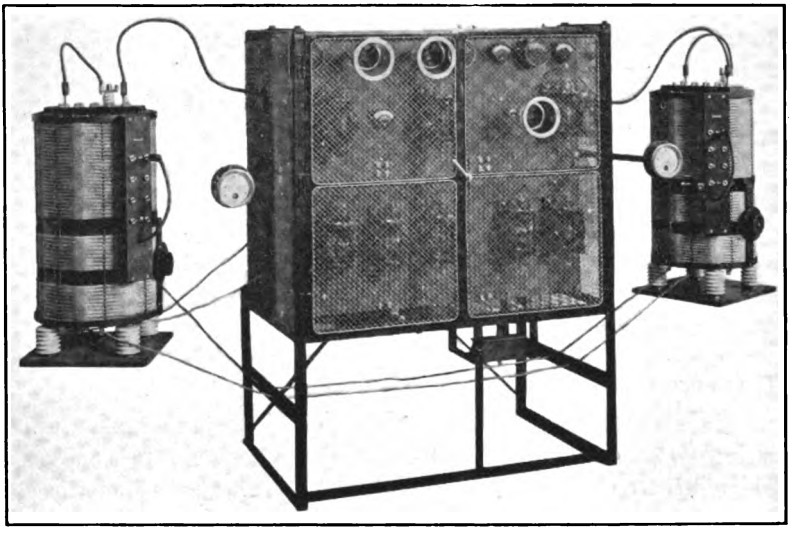

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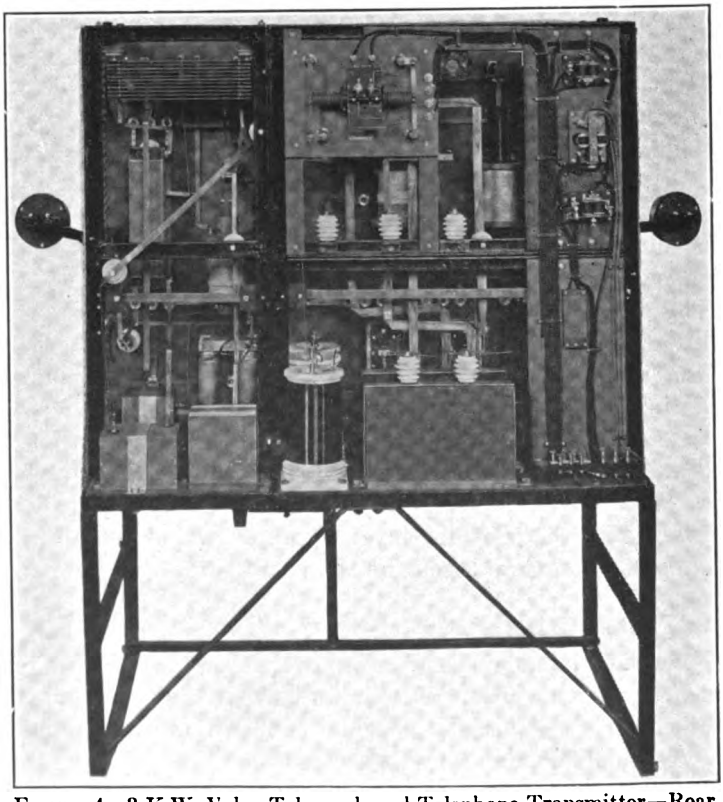

FIGURE 4—3 K.W. Valve Telegraph and Telephone Transmitter—Rear View Showing Control Relays

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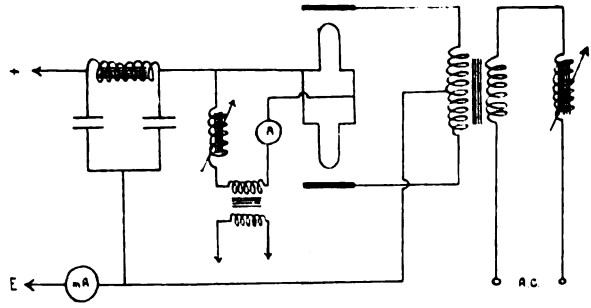

FIGURE 4A—Rectifier

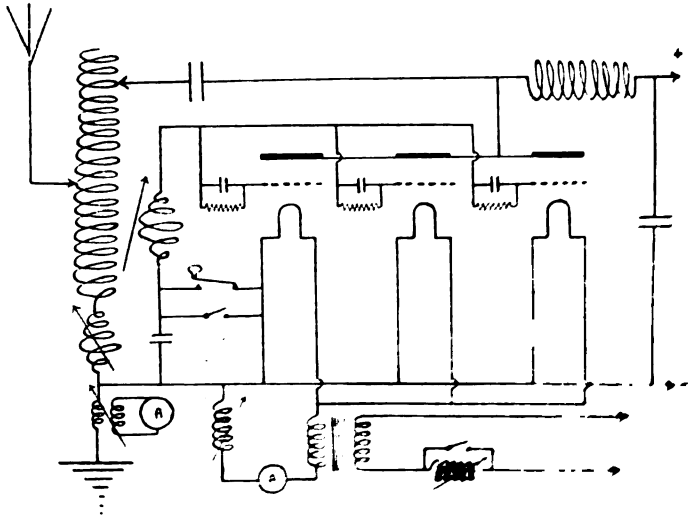

FIGURE 4B—Oscillator

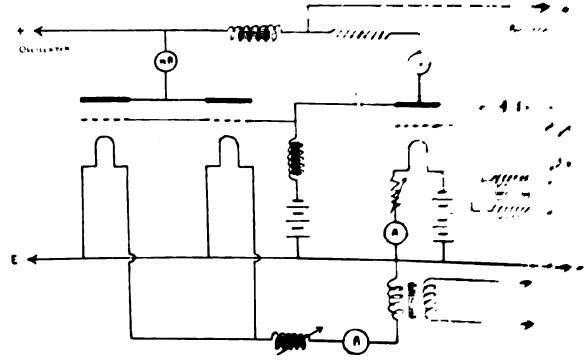

FIGURE 4C—Modulator

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supplied to a quenched spark gap of eight plates, thru suitable high frequency chokes. A mica condenser and flat spiral inductance form the primary closed circuit, which is coupled to the aerial circuit by another flat inductance, made of copper strip. The coupling is varied by altering the number of secondary turns, and the aerial is tuned by altering the tappings on other flat spiral inductances. By means of a switch which changes simultaneously the primary and secondary tappings, the wave length may instantly be changed to 450, 600, or 800 meters.

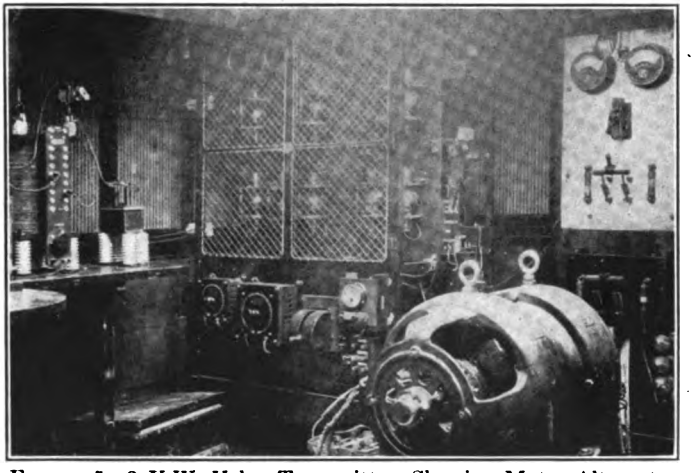

FIGURE 5—3 K.W. Valve Transmitter, Showing Motor Alternator and Aerial Inductance

The signaling key is placed in the primary circuit of the power transformer.

A hot wire ammeter is provided for measuring the aerial current, and a voltmeter for indicating the alternating current supply voltage.

Power to the transmitter is varied by altering the excitation of the alternator field, the number of plates in the spark gap being varied accordingly. Figure 6 is a view of the complete transmitter, and Figure 6A shows the connections.

EMERGENCY SPARK TRANSMITTER

For short distance working, and times when current from the yacht's supply is not available, an emergency spark transmitter is supplied. This derives its power from a storage battery of twelve cells, which supply current to a small inductor motor-alternator having an output of 0.25 kilowatt at 800 cycles fre-

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quency. This supply is stepped up to 2000 volts by a small transformer, and supplied to a quenched gap.

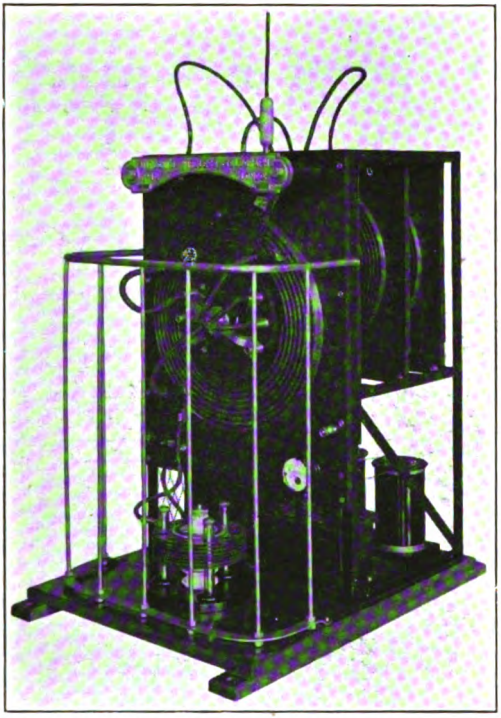

FIGURE 6—1.5 K.W. Quenched Gap Transmitter

This spark gap, a mica condenser, and a flat spiral inductance, form the closed circuit, which is directly coupled to the aerial circuit. Wave lengths of 300, 450, 600, and 800 meters may be obtained by a switch, part of which alters tappings in the closed circuit, and the other part alters tappings on the aerial inductance. Final tuning is effected by a variometer in the aerial circuit. resonance being indicated by an ammeter in series with the earth lead. The power is varied by adjustment of the generator field, and by inserting an iron-cored choke into the primary of the transformer. Signaling is carried out by means of a telegraphic key placed in the supply leads of the transformer. Figure 7 is a view of the transmitter, Figure 8 the motor alternator, and Figure 8A the connections of the set.

SHIP TUNER TYPE RECEIVER

This is a simple two circuit tuner in which the primary and

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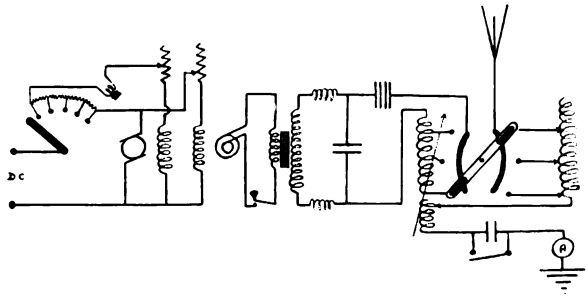

FIGURE 6A—1.5 K.W. Quenched Spark Transmitter

secondary windings are auto-coupled by means of a small coil provided with tappings. A wave range of from 200 to 25000 meters may be obtained by connecting various sized coils in the primary and secondary circuits by switches.

FIGURE 7—0.25 K W. Quenched Spark Transmitter

A continuous adjustment is obtained by variable condensers, that in the primary circuit being connected either in series or in parallel by means of a switch. A small fixed condenser may be also connected in parallel with the secondary condenser, in order to reach the maximum wave length. A coil of high impedance is connected across the antenna and earth terminals to by-pass any static charge from the aerial. The whole tuner is enclosed in a small copper-lined box, and is extremely compact for such

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a wide range of wave lengths. Figure 9 shows the tuner, and Figure 9A the connections.

Detector-Amplifier

Radio frequency amplification, rectification, and low frequency amplification are all carried out by one four-electrode valve. The air-cored radio frequency transformer is pro-

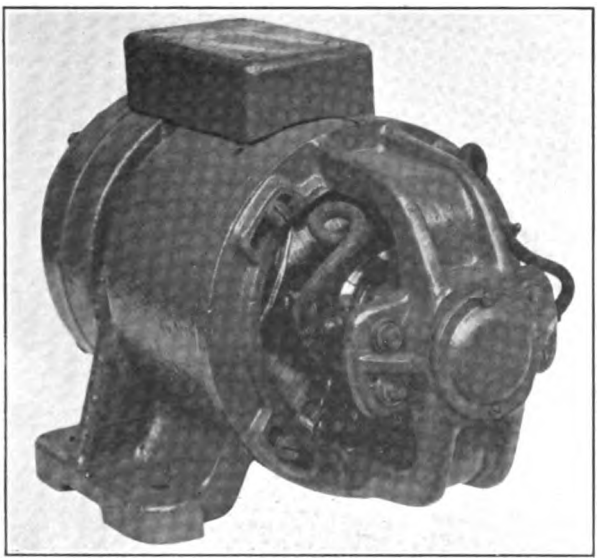

Figure 8—Motor Alternator for Emergency Transmitter

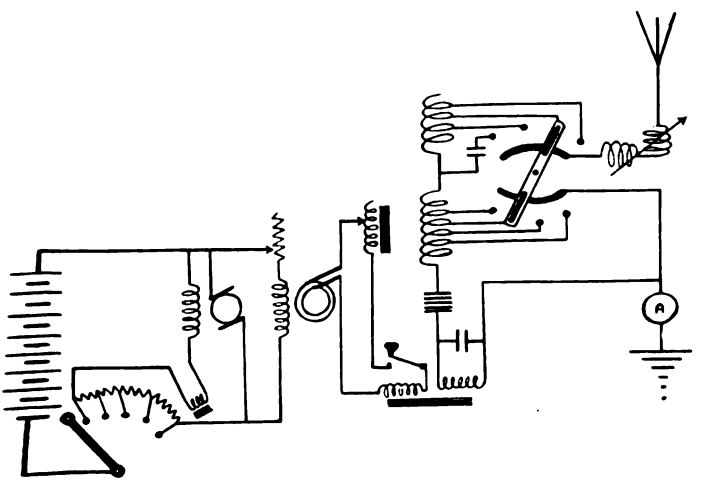

Figure 8A—0.25 K.W. Emergency Transmitter

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supplied to a quenched spark gap of eight plates, thru suitable high frequency chokes. A mica condenser and flat spiral inductance form the primary closed circuit, which is coupled to the aerial circuit by another flat inductance, made of copper strip. The coupling is varied by altering the number of secondary turns, and the aerial is tuned by altering the tappings on other flat spiral inductances. By means of a switch which changes simultaneously the primary and secondary tappings, the wave length may instantly be changed to 450, 600, or 800 meters.

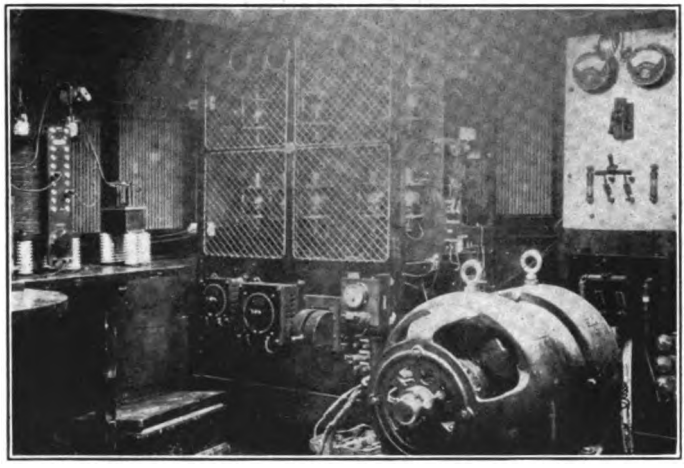

FIGURE 5—3 K.W. Valve Transmitter, Showing Motor Alternator and Aerial Inductance

The signaling key is placed in the primary circuit of the power transformer.

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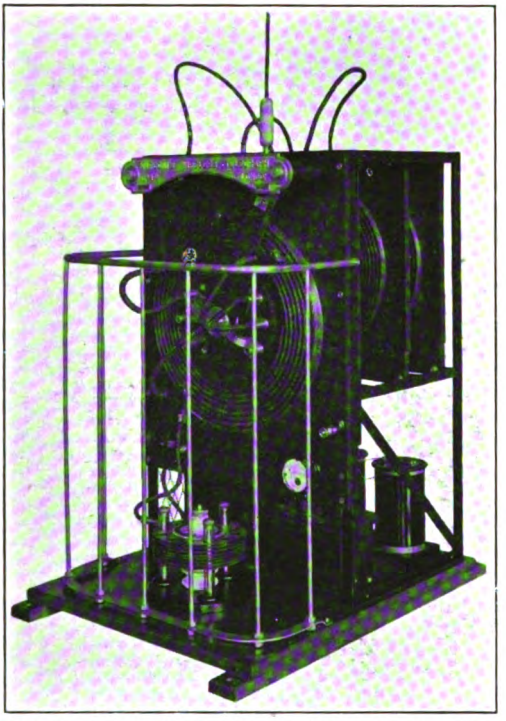

FIGURE C—1.5 K.W. Quenched Gap Transmitter

This spark gap, a mica condenser, and a flat spiral inductance, form the closed circuit, which is directly coupled to the aerial circuit. Wave lengths of 300, 450, 600, and 800 meters may be obtained by a switch, part of which alters tappings in the closed circuit, and the other part alters tappings on the aerial inductance. Final tuning is effected by a variometer in the aerial circuit. resonance being indicated by an ammeter in series with the earth lead. The power is varied by adjustment of the generator field, and by inserting an iron-cored choke into the primary of the transformer. Signaling is carried out by means of a telegraphic key placed in the supply leads of the transformer. Figure 7 is a view of the transmitter, Figure 8 the motor alternator, and Figure 8A the connections of the set.

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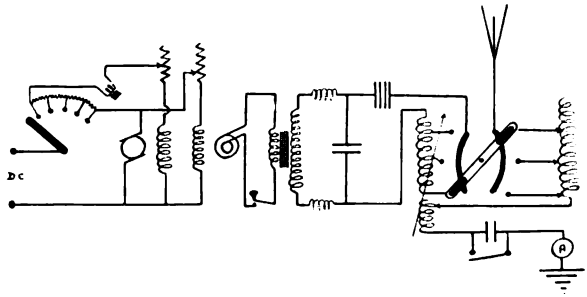

FIGURE 6A—1.5 K.W. Quenched Spark Transmitter

secondary windings are auto-coupled by means of a small coil provided with tappings. A wave range of from 200 to 25000 meters may be obtained by connecting various sized coils in the primary and secondary circuits by switches.

FIGURE 7—0.25 K W. Quenched Spark Transmitter

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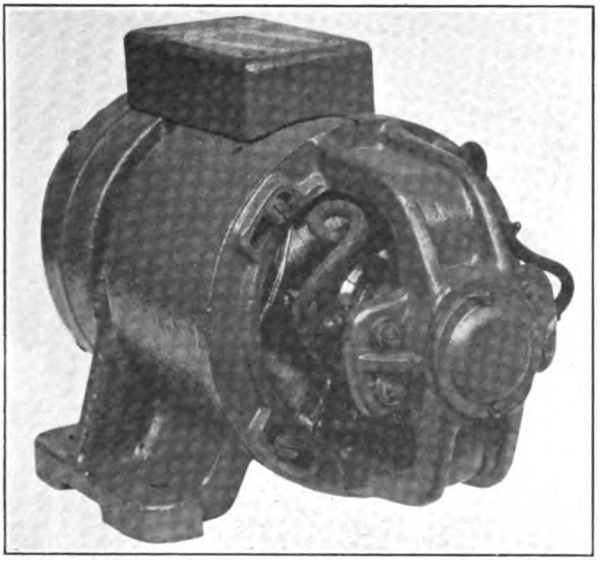

Figure 8—Motor Alternator for Emergency Transmitter

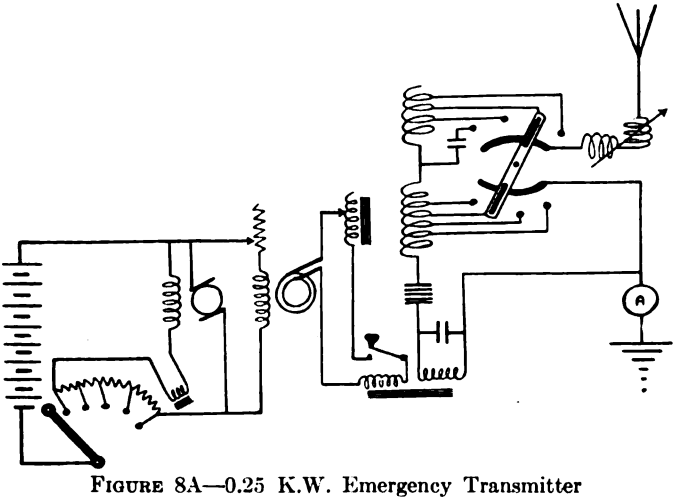

Figure 8A—0.25 K.W. Emergency Transmitter

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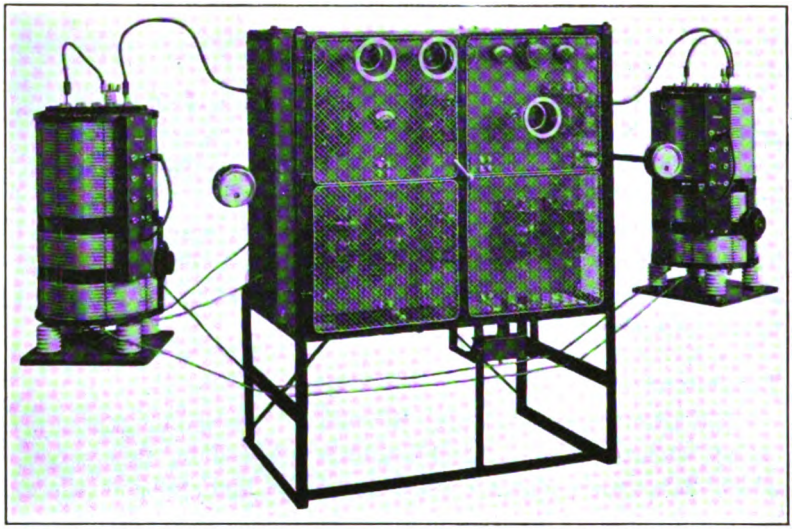

FIGURE 3—3 K.W. Valve Telegraph and Telephone Transmitter, Arranged for Inductive Coupling to Aerial

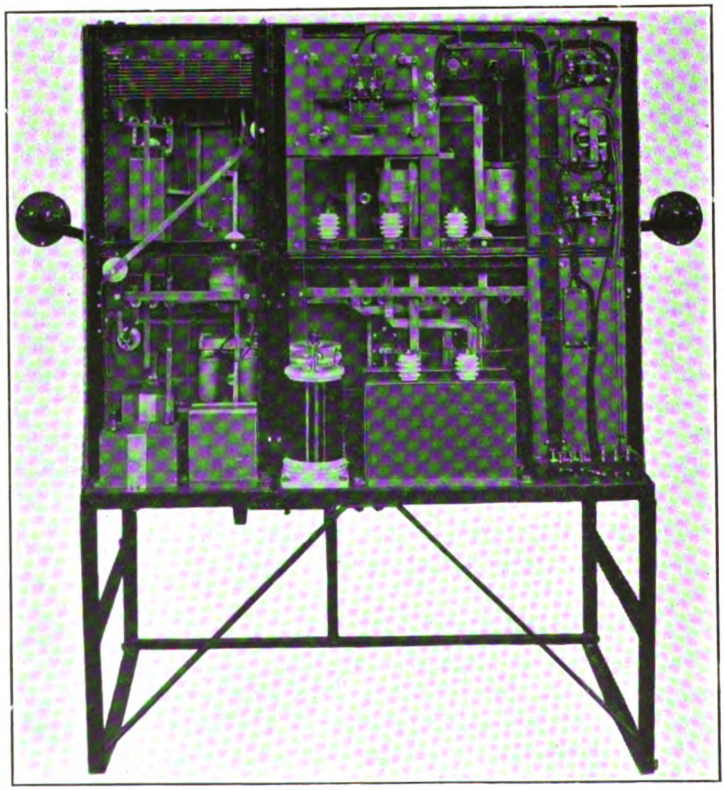

FIGURE 4—3 K.W. Valve Telegraph and Telephone Transmitter—Rear View Showing Control Relays

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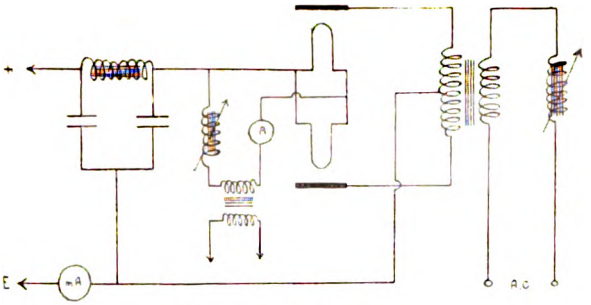

FIGURE 4A—Rectifier

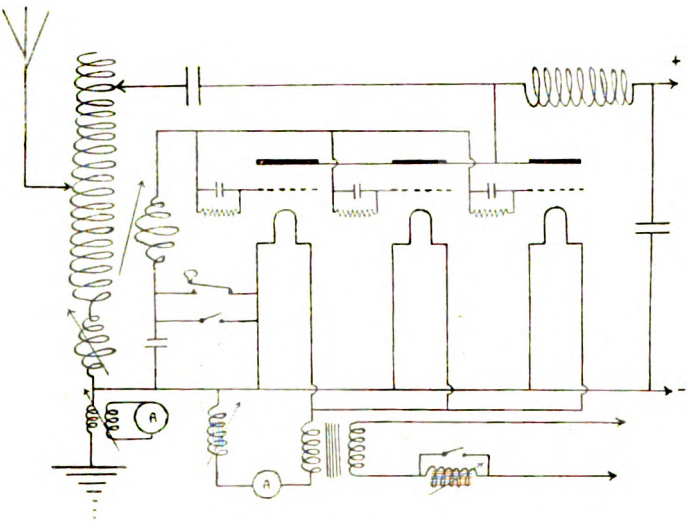

FIGURE 4B—Oscillator

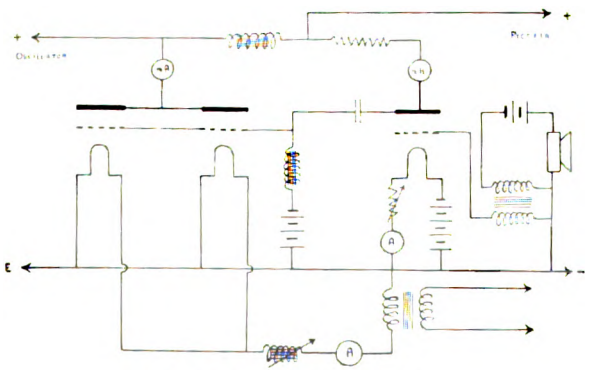

FIGURE 4C—Modulator

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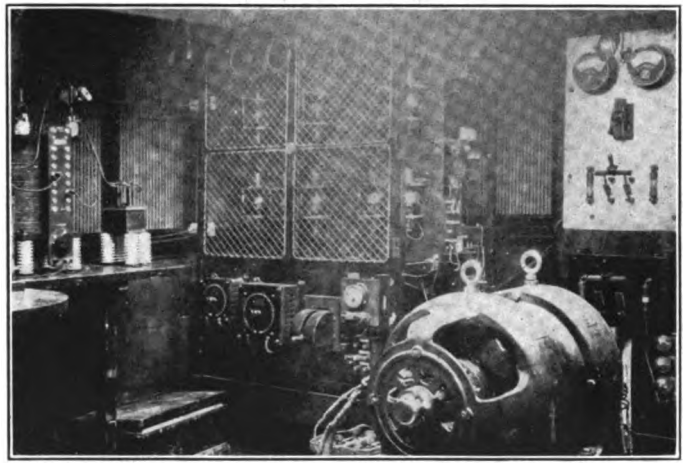
supplied to a quenched spark gap of eight plates, thru suitable high frequency chokes. A mica condenser and flat spiral inductance form the primary closed circuit, which is coupled to the aerial circuit by another flat inductance, made of copper strip. The coupling is varied by altering the number of secondary turns, and the aerial is tuned by altering the tappings on other flat spiral inductances. By means of a switch which changes simultaneously the primary and secondary tappings, the wave length may instantly be changed to 450, 600, or 800 meters.



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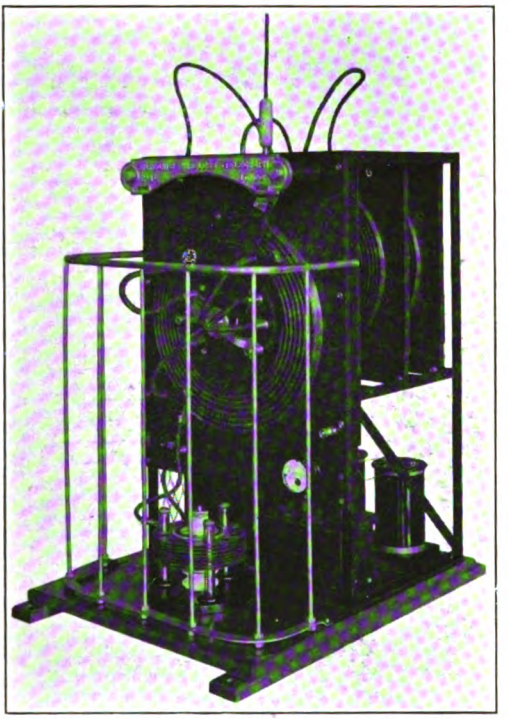

FIGURE C—1.5 K.W. Quenched Gap Transmitter

This spark gap, a mica condenser, and a flat spiral inductance, form the closed circuit, which is directly coupled to the aerial circuit. Wave lengths of 300, 450, 600, and 800 meters may be obtained by a switch, part of which alters tappings in the closed circuit, and the other part alters tappings on the aerial inductance. Final tuning is effected by a variometer in the aerial circuit, resonance being indicated by an ammeter in series with the earth lead. The power is varied by adjustment of the generator field, and by inserting an iron-cored choke into the primary of the transformer. Signaling is carried out by means of a telegraphic key placed in the supply leads of the transformer. Figure 7 is a view of the transmitter, Figure 8 the motor alternator, and Figure 8A the connections of the set.

SHIP TUNER TYPE RECEIVER

This is a simple two circuit tuner in which the primary and

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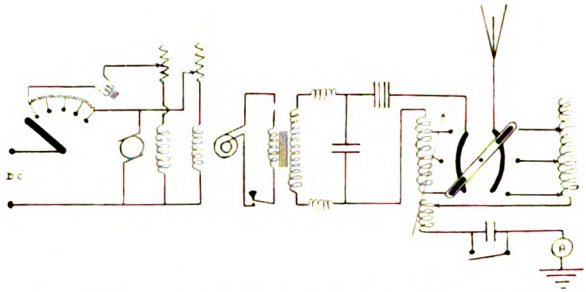

FIGURE 6A—1.5 K.W. Quenched Spark Transmitter

secondary windings are auto-coupled by means of a small coil provided with tappings. A wave range of from 200 to 25000 meters may be obtained by connecting various sized coils in the primary and secondary circuits by switches.

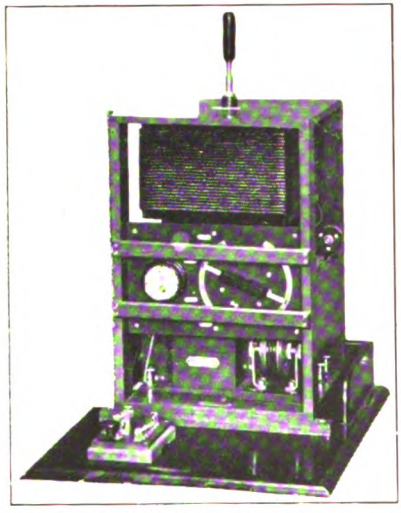

FIGURE 7—0.25 K.W. Quenched Spark Transmitter

A continuous adjustment is obtained by variable condensers, that in the primary circuit being connected either in series or in parallel by means of a switch. A small fixed condenser may be also connected in parallel with the secondary condenser, in order to reach the maximum wave length. A coil of high impedance is connected across the antenna and earth terminals to by-pass any static charge from the aerial. The whole tuner is enclosed in a small copper-lined box, and is extremely compact for such

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a wide range of wave lengths. Figure 9 shows the tuner, and Figure 9A the connections.

Detector-Amplifier

Radio frequency amplification, rectification, and low frequency amplification are all carried out by one four-electrode valve. The air-cored radio frequency transformer is pro-

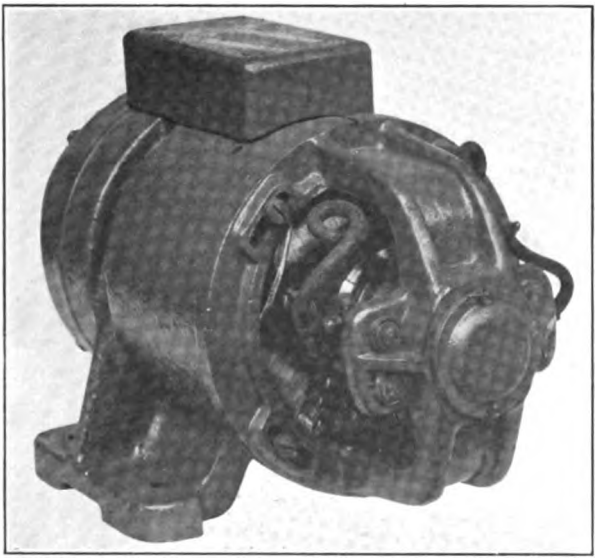

Figure 8—Motor Alternator for Emergency Transmitter

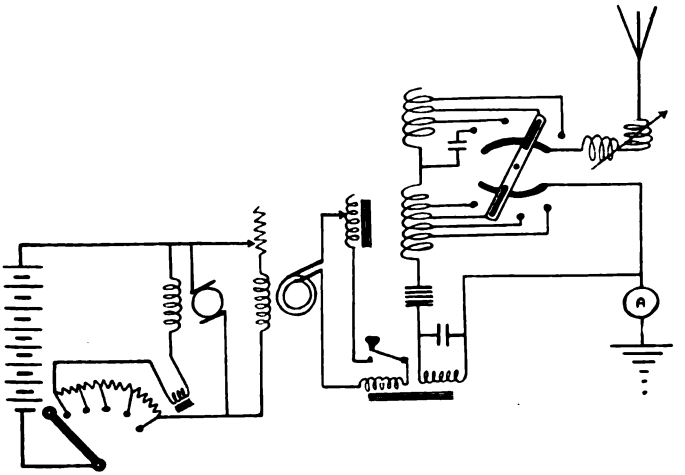

Figure 8A—0.25 K.W. Emergency Transmitter

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vided with three tappings, so that good amplification is obtained over the whole range of wave lengths. The signals from the tuner are applied to the inner grid circuit, and amplified into the outer grid circuit and primary of the air core transformer. From

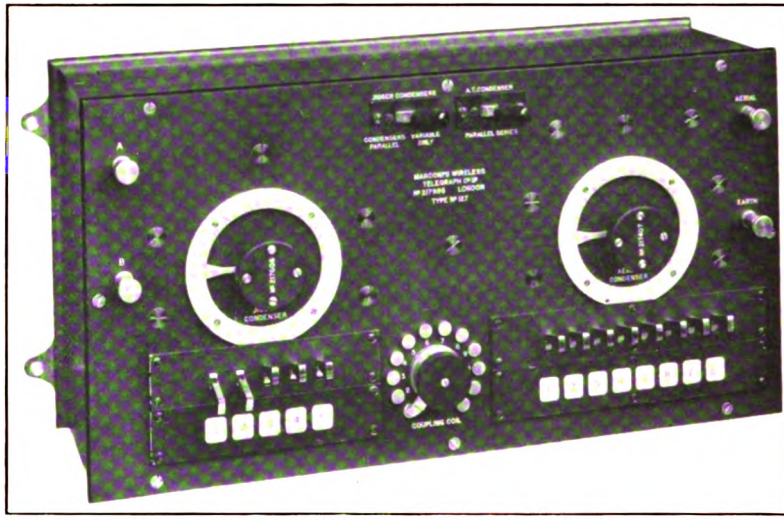

FIGURE 9—Receiving Tuner

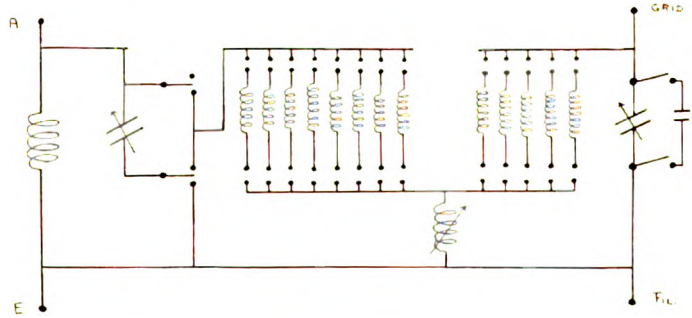

FIGURE 9A—Tuner

the secondary of this transformer they are rectified in the plate circuit, and transferred back to the inner grid circuit by way of the iron cored transformer. The amplified signals are then received by telephones in the outer grid circuit. Both the primary and secondary of the iron-cored transformer, and also the telephones, are shunted with condensers for by-passing the radio frequencies.

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A grid potentiometer and filament resistance are provided for controlling the point of working on the curve and the strength of signals. Figure 10 is a view of the amplifying detector, and Figure 10A shows the connections.

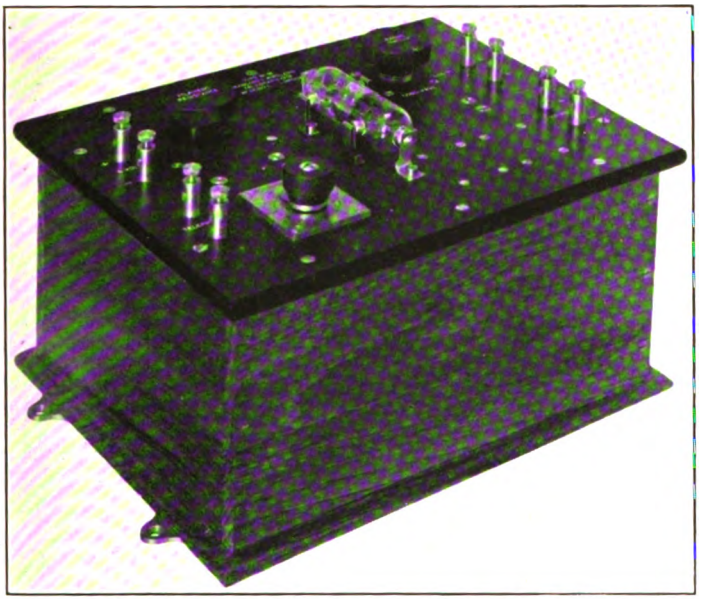

FIGURE 10—Amplifying Detector for Receiver

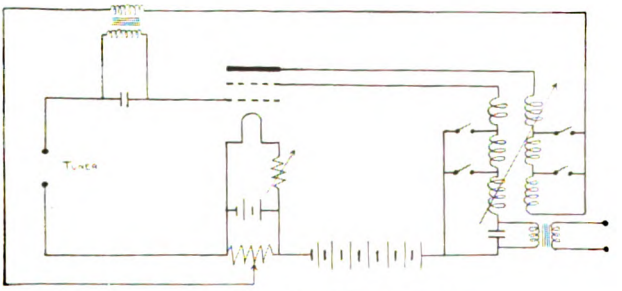

FIGURE 10A—Amplifier-Detector

LOCAL OSCILLATOR

For the reception of continuous waves, beat reception is employed. One valve is arranged to generate oscillations on all wave lengths from 200 to 25000 meters, by using flatly-wound coils tightly coupled together. Four pairs of coils cover the complete range of wave lengths, and continuous adjustment is ob-

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tained by a variable condenser, which may be connected across either the grid only, or across both coils. By this means, two wave length ranges are obtained for each pair of coils.

The coils are enclosed in an ebonite box and may be plugged in as required, each box containing two pairs and being reversed for changing the wave range. A small rotatable coupling coil is arranged in the vicinity of the coils in order to couple the oscillator to the receiver. Figure 11 is a view of the local oscillator, and Figure 11A gives the connections.

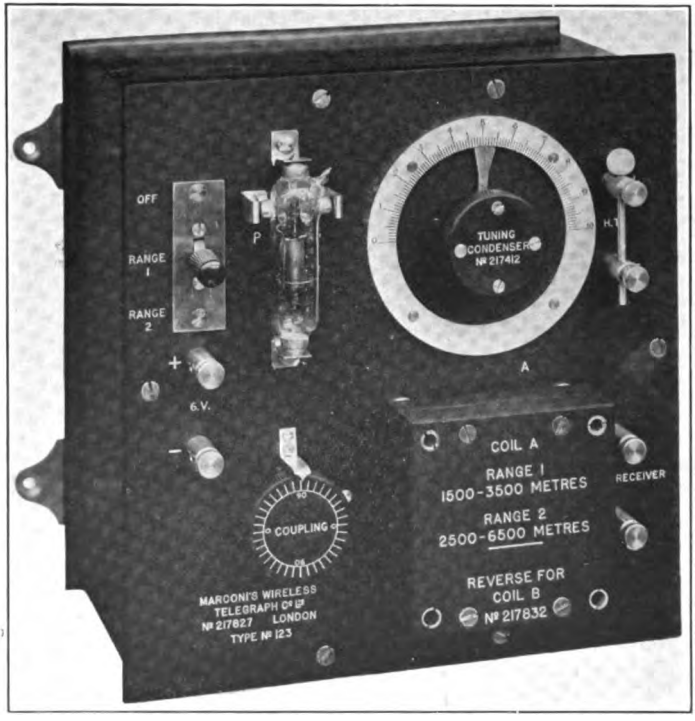

FIGURE 11—Local Oscillator for Receiver

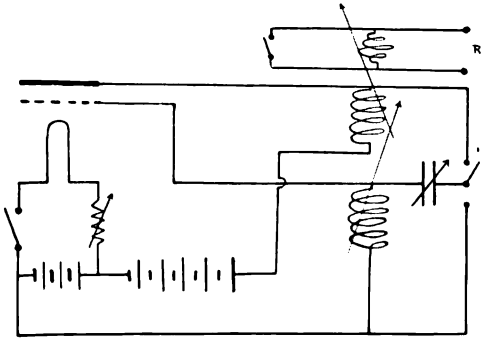

FIGURE 11A—Oscillator for Local Oscillator Reception

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The Radiogoniometer

Aerials

The special aerials for the direction finder are suspended over the bridge from a jumper stay running between the two masts. There are two triangular loops, one with its plane in the fore and aft line of the ship and the other athwartships. The four leads from these two loops are brought down thru lead-covered non-inductive cables into the chart room below, and led to the instrument, a view of which is shown in Figure 12. The antenna, direction finder, and tuner are shown in Figure 12A.

The width of each loop is 24 feet (7.87 m.) and the height is 34 feet (11.15 m.).

Goniometer

The goniometer is wound with two field coils at right angles to one another, one being connected to the fore and aft loops and

Figure 12—The Direction Finder

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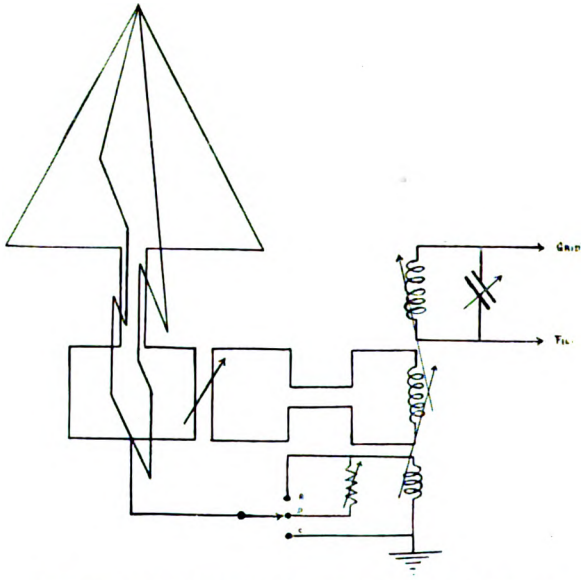

FIGURE 12A—Direction Finder Aerial and Circuits

the other to the athwartships loop; these are plainly shown in Figure 13, which is a view of the goniometer unit.

Within these two field coils a rotatable search coil is placed, the spindle of which is attached to a pointer which moves over a circular scale, graduated in degrees from 0 to 360.

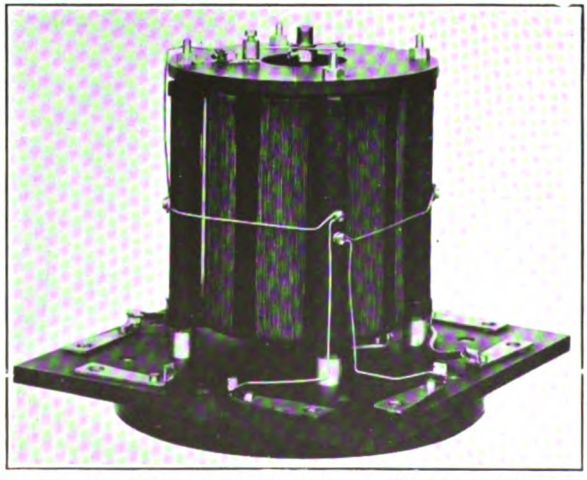

FIGURE 13—View of the Goniometer, Showing Field Coils

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Tuner

The search coil is tightly coupled thru a transformer to the amplifier, the secondary of this transformer being tuned by a variable condenser. A tertiary winding on this transformer is connected from the centre points of the field coils, and thru a damping resistance to earth.

A range of from 300-4000 meters may be obtained by using three differently wound transformers, any of which may be brought into operation by means of a switch.

Amplifier-Detector

The radio frequency amplifier employs four valves. These are coupled together by air cored transformers, which are wound with high resistance wire so as to have a fairly flat resonance curve and give good amplification over the entire range of wave lengths used. The potential of the grids is adjusted by a potentiometer and the filament brilliancy by a variable resistance.

One valve is used as a rectifier and one as a low frequency amplifier, these being coupled by iron-cored transformers.

The low frequency amplifier may be cut out by a switch if the signals are too loud. Figure 14 is a rear view of the amplifier, and Figure 14A shows the connections.

General

A relay is fitted which connects the four aerial leads to earth when the instrument is not in use. By operating a switch, the instrument may be used as an ordinary receiver, using the whole aerial system as a vertical aerial. This gives a uniform polar diagram and is useful for standing by. For direction finding, the effect of the two loops is used, this giving a figure-of-eight polar diagram.

For finding the actual sense of the direction, the effect of the two loops is combined with that of the vertical aerial system, thus giving a cardioid polar diagram.

Either of these methods of reception may be employed by the operation of the single switch shown in the diagram, Figure 12A. The switch at position *A* gives a uniform polar curve; at *B* a cardioid, and at *C* a figure-of-eight diagram.

Operation

To take bearings with the instrument, the station, the bearing of which it is desired to find, is first picked up, using the stand-by position, and tuning on the variable condenser. As

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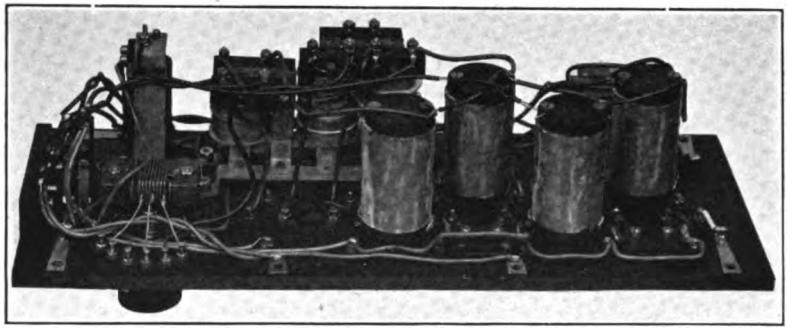

FIGURE 14—Amplifier for Direction Finder, Rear View

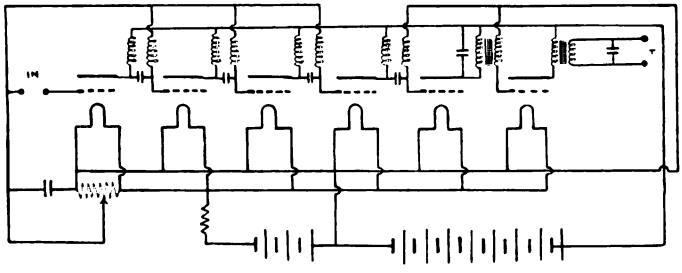

FIGURE 14A—Amplifier-Detector

soon as the signals are heard, they are rapidly tuned in and then the switch is put over to the position for using the two loops. The strength of the signals is adjusted to a suitable value by altering the filament brilliancy or by cutting out the low frequency amplifier, and the pointer of the goniometer is swung around the position where signals are a minimum strength, and two points at which they are of a similar strength are noted; the mean of these two points is taken as the bearing required.

If possible, several sets of readings are obtained, and an average taken. Altho the sense of the direction is usually obvious, this may always be checked by switching on the cardioid polar diagram, the sense then being indicated by a small pointer attached at right angles to the other pointer. From this description it sounds as if the taking of a bearing were a lengthy process, but actually, it is a matter of seconds.

Any known error of the instrument is, of course, allowed for before the reading is given to the navigating officer. Figure 15 is a view of the direction finder situated in the chart room of the yacht.

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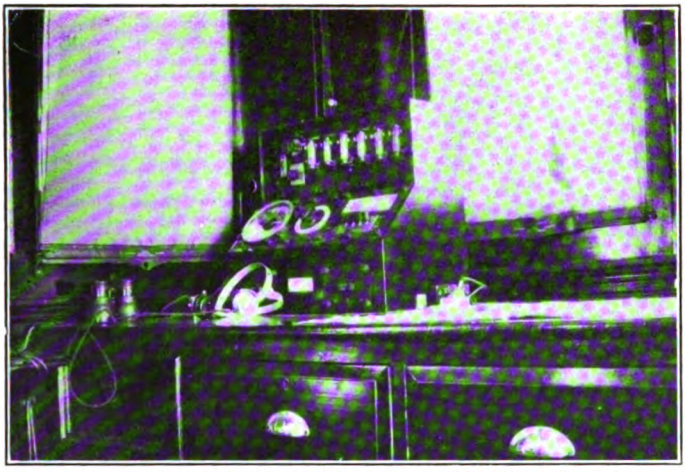

FIGURE 15—Direction Finder Located in Chart Room

ACKNOWLEDGMENTS

The author is indebted to Senatore Marconi for permission to publish these particulars of his yacht and its radio equipment, and to Marconi's Wireless Telegraph Company, Ltd., for supplying photographs of some of the apparatus.

SUMMARY: A description is given of the continuous wave telegraph, telephone, and main and emergency spark transmitters, and of the receivers and radio goniometer of the steam yacht "Elettra."

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THE DEVELOPMENT OF THE STANDARD DESIGN FOR SELF-SUPPORTING RADIO TOWERS FOR THE UNITED FRUIT AND TROPICAL RADIO TELEGRAPH COMPANIES*

By

Albert W. Buel

(Consulting Engineer, United Fruit Company, Radio Department, New York)

This paper is largely collated from records of designs, construction, maintenance, reports on failures, reconstruction and reinforcement of some 20 or 30 radio towers, built by the United Fruit Company and the Tropical Radio Telegraph Company, between 1906 and 1923, and some other designs for towers from 100 ft (30.5 m.) to 1,000 ft. (305 m.) high. It has been prepared at the request of Mr. George S. Davis, General Manager of the Radio Department of the United Fruit Company and of the Tropical Radio Telegraph Company, to whom the author is indebted for much assistance, advice, and consultation, and for the photographs used in illustrations. All of this work since 1911 has been under the personal direction of Mr. Davis, except that the original design of the 200 ft. (61 m.) towers made in 1906 and the 50 ft. (15 m.) extension added in 1910 came under his predecessor, Mr. M. Musgrave. To Mr. Davis is due the credit for the adoption of the three-legged or triangular tower and for a number of important improvements in details.

The Problem and Suggestive Precedents

In June, 1906, Mr. Mack Musgrave, then Superintendent of the Fruit Company's telegraph service, brought his radio tower problem to the author, who had designed bridges and other structures for that Company. He wanted 200 ft. (61 m.) towers

*Received by the Editor, May 3, 1923.

For the information of our readers, the following articles on towers and masts for radio antenna supports have appeared in the Proceedings of The Institute of Radio Engineers:

R. A. Weagant, "Design and Construction of Guy-Supported Towers for Radio Telegraphy," volume 3, number 2, June, 1915, page 135.

C. F. Elwell, "Wooden Lattice Masts," volume 3, number 2, June, 1915, page 161.

C. F. Elwell, "The Rome Radio Station of the Italian Navy," volume 8, number 3, June, 1920, pages 193-196.

In conjunction with Mr. Buel's paper, these should be of material assistance to radio station designer.—Editor.

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to carry 20 antenna wires equivalent to number 10 copper wire[1] in size, with a span of 300 ft. (91.5 m.) between 42 ft. (12.6 m.) yard arms. He had a fabricator's design and estimate for 200 ft. (61 m.) towers, 14 ft. (4.6 m.) square at base and at top—rectangular in elevation—but the weights and cost were much higher than they were then willing to accept.

There were not then many precedents for steel towers, outside of viaduct towers (of which the author had recently designed some 230 ft. (70 m.) high), and of the Eiffel Tower in Paris. It was obvious that the curved outline of the legs of the latter tends to reduce the weight of the bracing—(horizontal struts and diagonals)—and also of the surface area exposed to wind.

In his discussion of a paper on the Kinsua Viaduct on the Erie Railroad, the towers of which were designed without diagonals, Mr. Gustave Lindenthal suggested the use of rod diagonals to reduce the bending stresses in the legs and struts without sacrificing the advantages of the design under discussion. ("Transactions of the American Society of Civil Engineers," volume XLVI.) This suggestion had been adopted by the author for the 230 ft. (70 m.) towers of the Las Vacas Viaduct in Guatemala with considerable success, demonstrating the advantages of adjustable rod bracing in special cases, notwithstanding its practical abandonment by most structural steel designers for important structures for many years past.

With these points in mind, it will be readily seen how the outline of the original design for the 200 ft. (61 m.) tower quickly took form, with its curved legs and rod diagonals. Figure 1 shows the completed design from which ten or more towers were built, some of them with the 50 ft. (15 m.) extension, also shown in the figure. This was added in 1910 without increasing any of the sections below and without consulting the designer, who did not know of the increase in height to 250 ft. (76 m.) until after one of the towers failed in the hurricane of 1914. Some of these towers, extended to 250 ft. (76 m.) were used in a four-tower station, 300 by 600 ft. (91.5 by 183 m.) center to center of towers, with messenger cables in the 300 ft. (91.5 m.) spans and the antenna wires on the 600 ft. (183 m.) span. This increased the wire pull from the design load of 2,000 lbs. (920 kg.) to 7,000 lbs. (3,220 kg.), and the resulting moment at the base from 400,000 ft. lbs. (56,000 kg. m.) to 1,750,000 ft. lbs. (240,000 kg. m.), or 437.5 percent. Still these towers stood up until a really violent hurricane came along in 1914. Some of the first 200 ft. (61 m.)

[1] Diameter of number 10 wire = 0.102 inch = 0.259 cm.

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towers are still in service, and it was their excellent behavior in high winds that was, no doubt, responsible for the increase in height and wire pull without increase in sections. Figure 2 shows two of these towers at Puerto Limon, Costa Rica, and Figure 3 the tower at Bocas del Toro, Panama.

As might be expected from an examination of the design, the towers are very flexible. We have come to recognize this as a desirable feature, but it was not appreciated at the time. The effects of flexibility will be referred to in another place.

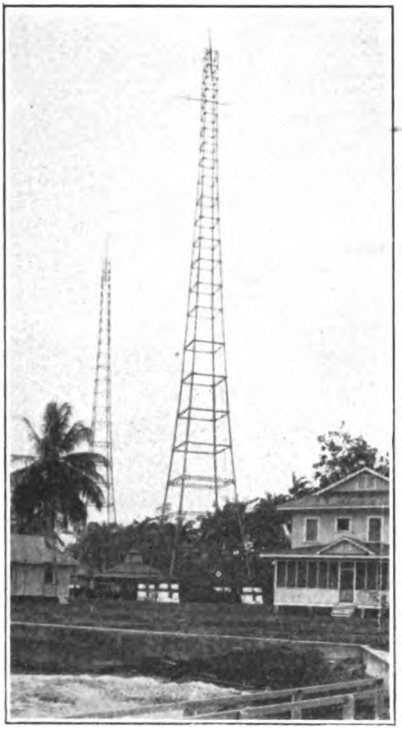

FIGURE 2—General View of the United Fruit Company Station, Limon, Costa Rica, 1923

A story has been told about one of the original 200 ft. (61 m.) stations, that the flexure of the towers was so considerable as to cause alarm, to the extent that steel guys were put on. One day a hurricane came and the first thing it did was to break the guys but the towers stood. Then it carried away the station buildings and the crew went under the towers for protection.

The 1906 design was made to keep the cost as low as possible, and this controlled shop details as well as weights, of course

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limited by considerations of safety. This will explain the use of the economical but not entirely satisfactory hitch of the diagonal rods. The first towers were insulated at the base of the legs and were all galvanized, but both of these features have been omitted in later designs. The records of costs of maintenance covering a period of fifteen years show that frequent painting is cheaper than galvanizing. Moreover, 1/16 inch (0.16 cm.) of steel is about as good for protection as the thin film of zinc. In all recent designs the minimum thickness of metal is 5/16 inch (0.79 cm.) for locations in the tropics or near the sea coast, and ¼ inch (0.63 cm.) for the interior in dry

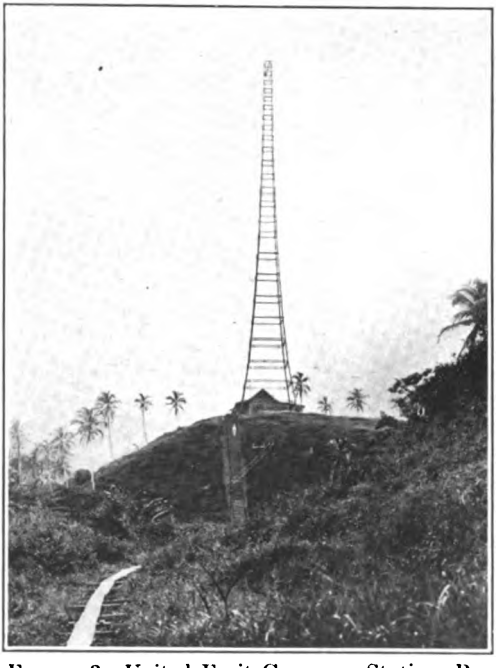

FIGURE 3—United Fruit Company Station, Bocas del Toro, Panama, 1904. First Radio Station to be erected in Central America

climates. All these towers have bolted field connections with ordinary rough bolts, not turned, which have proven satisfactory. They are inspected annually and repainted when required —on an average, about once every two years.

The original towers weighed 22.5 tons (21,000 kg.) each, and with the 50 ft. (15 m.) extension, added in 1910, 26 tons (24,200 kg.).

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Three Hurricane Wrecks and Resulting Design Improvements

In 1914, Swan Island was swept by a hurricane which wrecked the upper part of tower number 4 (Figure 4), the lower 75 ft. (22.8 m.) remaining standing and not damaged. The other three towers were not injured. The reports of this wreck were not in sufficient detail to give a very definite idea of the intensity or character of the force that wrecked the tower. As three of

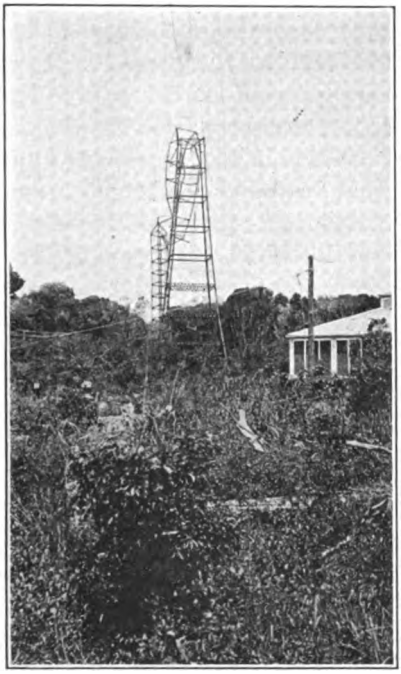

Figure 4—Tower, United Fruit Company, Swan Island Station, blown down in hurricane, June 6, 1914

the towers and nearly half of the weight of steel in the fourth remained in place and serviceable, and as the excessive cost of general reinforcing in the field had to be considered, together with the fact that we really had no reliable data on what wind force would have to be met in a tropical hurricane, it was decided not to attempt general reinforcement to resist hurricane force, but to make the towers good for the ordinary high wind of 30 lbs. per square foot (14.7 kg. per sq. m.), or about 90 miles per hour (144 km. per hr.), and for the 7,000 lbs. (3,220 kg.) pull of the messenger cables at a height of 250 feet (76 m.).

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In computing wire pull and sag under wind loads, we take into account the elastic stretch of the wires and cables and deflections of the towers. Some consideration of this problem will make it apparent that these elastic changes have a remarkable influence on the wind increment in wire pull, especially where messenger cable systems are used. In some cases only 17 percent of the wind load on wires acted to stress the towers, the other 83 percent being compensated by deflection of towers, stretch of wires and cables, and consequent increased sag. This taught us two things: first, the advantage of flexibility in radio towers, which is increased by the curved legs; and, second, that the sag of mat under no wind can be so fixed that at maximum wind the pull will not exceed a fixed figure, at the same time allowing a greater average height or less sag under no wind.

It is probable that flexibility was largely responsible for these towers standing up as long as they did, under wire-pull moments over four times those for which they were designed.

Retaining the 50 ft. (15 m.) extension added in 1910, and the messenger cable pull of 7,000 lbs. (3,220 kg.), there was no economical way to avoid the use of guys; but by hitching the guys at a point as far below the top as the strength of the members permitted, we retained considerable flexibility and reduced the length of the guys. This point worked out at "U", Figure 5, 230 ft. (70 m.) above the base.The guys were run at an angle of about 30 degrees from the vertical, down to concrete anchors in the ground, and were placed only on the outside of the mat at an angle of 90 degrees to each other in plan, so as to resist the resultant pull of antenna wires and messenger cable, as shown in Figure 5. (Guys in the quadrant under the mat were not permitted.)

The leg sections P-T were deficient in strength and were made heavier for the new sections of tower number 4. In the other three towers new horizontal and "K" bracing was introduced between P and Q, as indicated in Figure 5, to reinforce these members by reducing the unsupported length of the columns.

With the towers thus guyed and reinforced the station (Figure 6) remained in service until an exceptionally severe hurricane swept Swan Island in 1916. Even then the towers stood until the antenna wires went out and an open hook of a messenger cable turnbuckle opened up, leaving the tower without support or assistance from the guys where the latter were to leeward.

The estimated velocity of the wind in this hurricane, according to the most reliable reports, was 130 miles per hour (208 km.

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per hr.), and computations of the ultimate strength of the leg sections in which initial failure took place, indicated a velocity of 137 miles per hour (219 km. per hr.). For these computations Tetmajer's formula, $f = 44{,}000 - 162.l/r$, was used. This formula has proven most reliable for columns of this particular type; some tests planned by the author and made at the Allegheny Arsenal station of the Bureau of Standards are given on page 600, volume LXXVI, "Transactions of the American Society of Civil Engineers," showing an agreement between computed and actual test loads closer than four percent, and closer than one percent, if totals of all tests are compared. If secondary stresses were not all given sufficient weight, the error would result in too high a figure for the indicated velocity, therefore

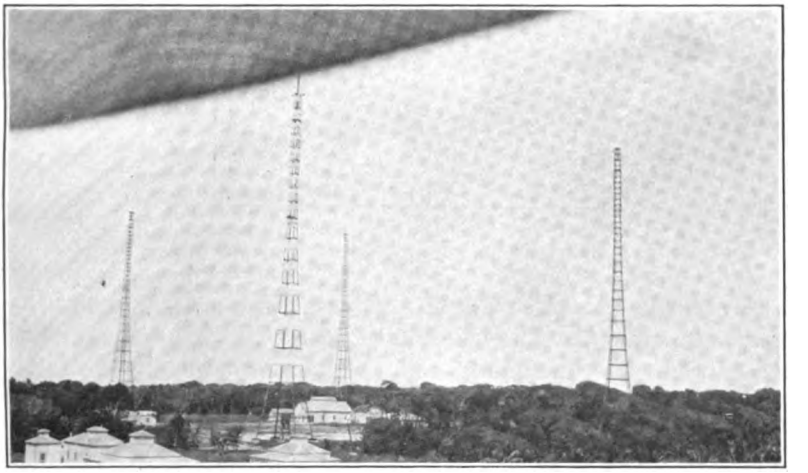

FIGURE 6—United Fruit Company, Swan Island Station before 1916 hurricane and as re-erected

the true velocity may have been somewhat less than 137 miles per hour (219 km. per hr.), but could not have materially exceeded that figure.

This storm mowed down a cocoanut palm grove 25 years old, with trees 18 inches in diameter—fairly conclusive evidence that its intensity had not been exceeded during that time. Figures 7 and 8 show this grove before and after the storm.

The hurricane started early in the morning, out of the North or Northeast. Between 8.00 and 8.50 A. M. it had gotten around to East by North, when the antenna wires broke loose from the messenger cable between towers 1 and 4, and an open hook of

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FIGURE 7—View in Cocoanut Grove, Swan Island, prior to 1916 hurricane

a turnbuckle on this cable opened up, leaving tower number 1 without support. The wind being nearly East, tower number 1 collapsed. Towers 2 and 3 collapsed several hours later, when

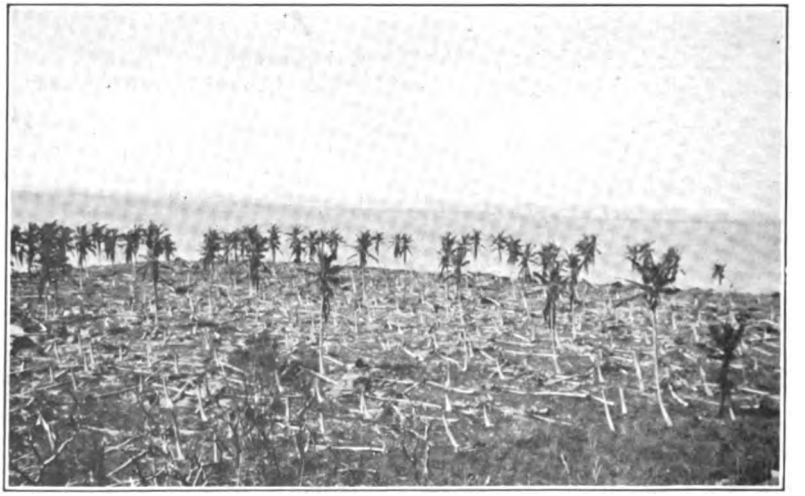

FIGURE 8—Cocoanut Grove after hurricane (1916)

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the wind had swung around into the Southeast quarter. The lay of the wreckage of these two towers indicates a wind from the South-Southeast, but if the messenger cable was still in, it may have anywhere in the Southeast quarter.

Tower number 4 remained standing. Its upper stories had been rebuilt the previous year with a few members made heavier, but it was not constructed to stand hurricane force, nor even a 90-mile-per-hour (144 km. per hr.) blow without support of the guys and messenger cable. It was evidently saved by the fact that the wind did not get around into a quarter where the guys could not lend it their support. If the turnbuckle hook had not opened up, possibly tower number 1 would not have been wrecked.

Figure 9 shows clearly the column buckle in story *Q-R*, just above the reinforcing "*K*" braces added in 1914-15. The bends at panel point *H*, Figures 10 and 11, are plainly due to the pull of the falling upper sections and in no way resemble the buckling caused by column flexure.

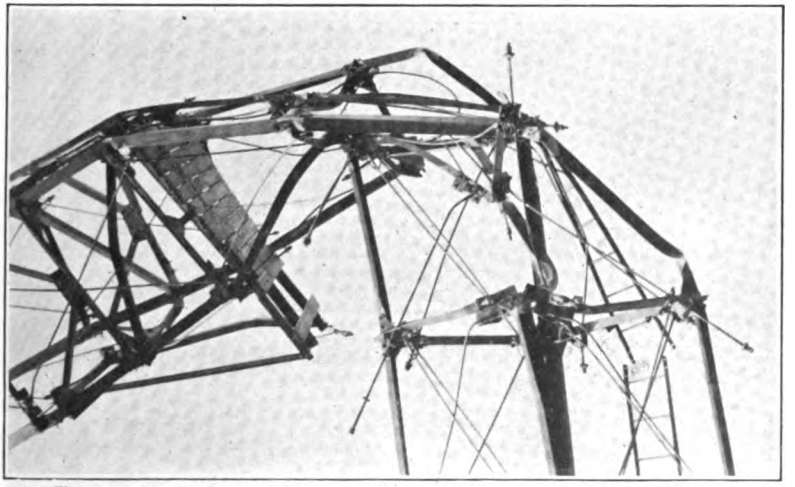

FIGURE 9—Tower 1, between "O" and "S" after hurricane of 1916

Figures 12, 13 and 14, show that towers 2 and 3 failed first by column buckling in story *O-P*, just below the story reinforced with "*K*" braces.

Figure 15 shows how the falling tops tore and stripped the lower stories from *P* to *H*. In no place is there any indication of column failure or tension failure between *P* and *H*, except the tearing clearly due to the stripping effect of the falling tops.

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The sequence of the occurrence during the collapse may be closely read from the illustrations as follows:

1. The failure of messenger cable between towers 1 and 4, the wind being nearly East, depriving tower 1 of support from the East guy of tower 4, the leeward legs in story *Q-R* buckled.

2. The top sections above *Q*, falling, pulled down the sections from *O* to *H*, this stripping effect being arrested when the top struck the ground.

3. The cushion effect due to the crushing of section *T-Z* against the ground saved the lower sections, *P* to *T* from total demolition.

4. Several hours later, when the wind had veered towards the South, towers 2 and 3 failed in exactly the same way except that legs buckled in story *O-P* instead of in story *Q-R*. With the wind in this quarter the guys were not effective. Until the wind got around into this direction the guys held the towers safely, the messenger cable co-operating to support tower 2.

In all these cases the primary failure occurred in stories adjoining that in which reinforcing "*K*" braces were inserted

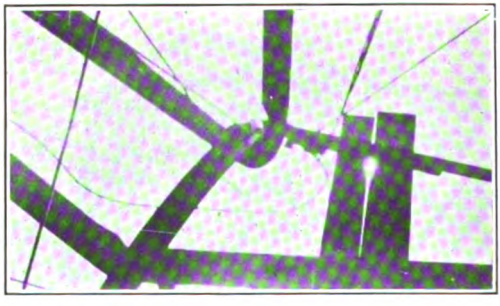

FIGURE 10—Tower 1, at "H" after 1916 wreck

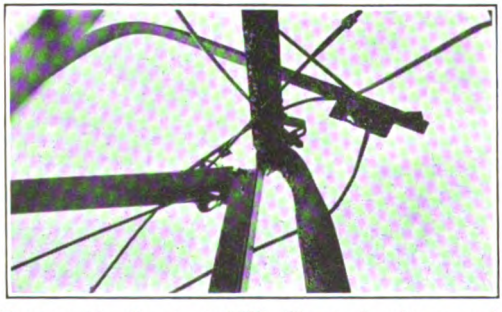

FIGURE 11—Strut at "H", Tower 1, after wreck

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in 1914-15, to strengthen this critical point. Had these not been added the initial failure would, undoubtedly, have occurred in leg sections *P-Q* in all cases.

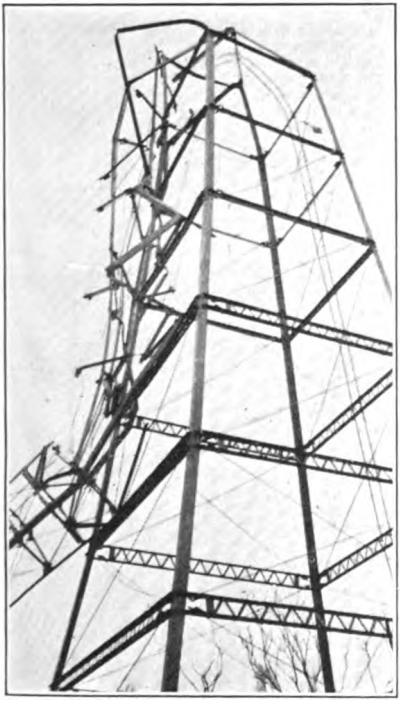

FIGURE 12—Tower 2, fallen part of tower from 'G" to "R"

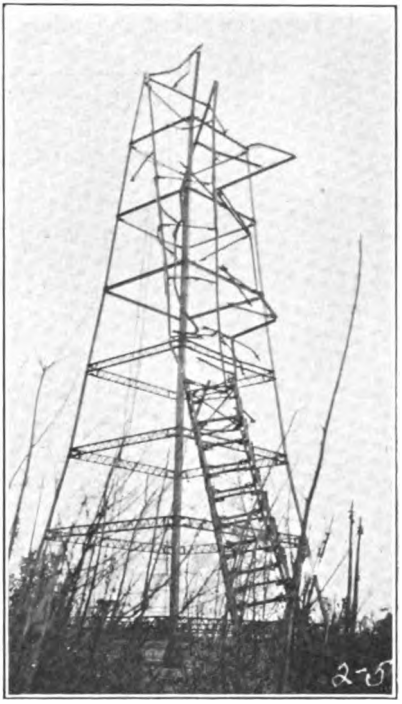

FIGURE 13—Tower 2, wrecked part from "G" to "Q"

A study of the stresses and ultimate strength of the members indicates that the towers failed under an average wind load of about 50 lbs. per sq. ft. (245 kg. per sq. m.). Taking Sir Benjamin Baker's ratio of average pressure on large areas—2/3 of the maximum shown at the same time on small gauges—as shown by the Forth Bridge experiments, the maximum intensity at Swan Island in 1916 probably reached 75 lbs. per sq. ft. (368 kg. per sq. m.) for small areas. In high velocities, according to several authorities, wind is more gusty. Therefore the average pressure on large areas would naturally be expected to be a smaller percentage of the maximum on small areas in high than in light winds. The Forth Bridge experiments showed this to be so for the range of velocities observed, up to 96 miles per hour (154 km. per hr.), and there can be little room for doubt

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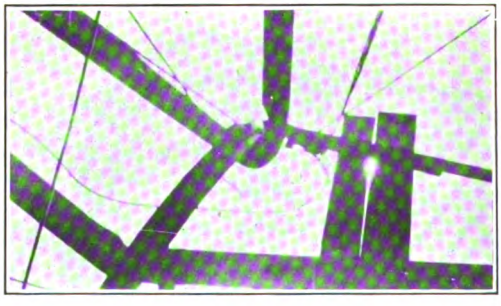

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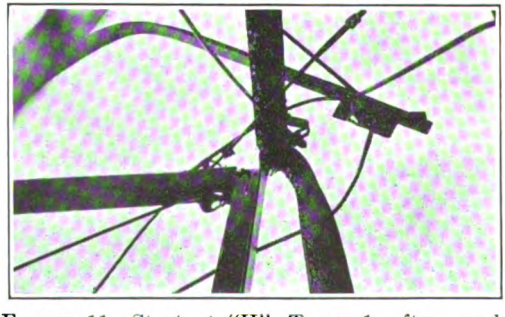

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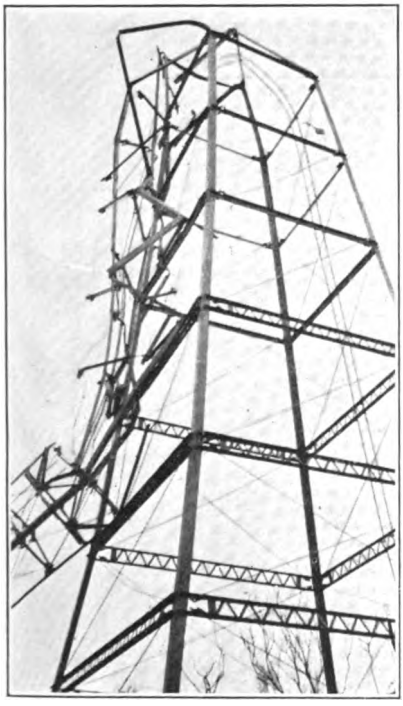
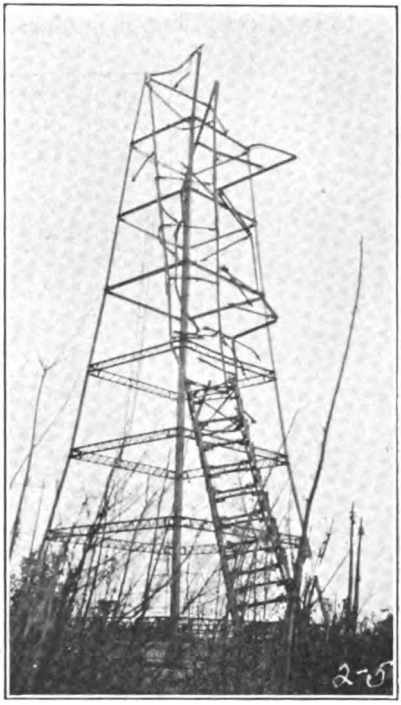


FIGURE 12—Tower 2, fallen part of tower from "G" to "R"



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that it is still more true for velocities between 120 and 140 miles per hour (192 and 224 km. per hr.). The velocity corresponding to 75 lbs. per sq. ft. (368 kg. per sq. m.) is, by our formula, 137 miles per hour (220 km. per hr.). This estimate is not claimed to be precise, but it is about the most reliable as yet available.

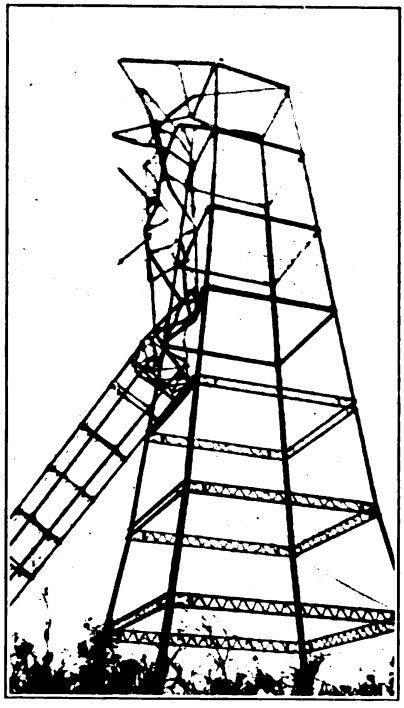

FIGURE 14—Tower 3, showing wreckage from point "H" to point "S"

At any rate, and whether this velocity and the factor of 2/3 for average pressures are correct or not, we can put considerable confidence in the deduction of 50 lbs. per sq. ft. (245 kg. per sq. m.) average, on the area involved at the moment of failure, and that is the value for which we have most need.

If that is true, it is not necessary to know the maximum velocity and intensity of pressure on small areas, but if their relative values were known we might determine the true law of these variations, which would be a great aid in specifying wind loads for different conditions and heights of towers. However, these deductions were considered a sufficient basis to warrant an attempt to build or reconstruct radio towers to be hurricane-proof.

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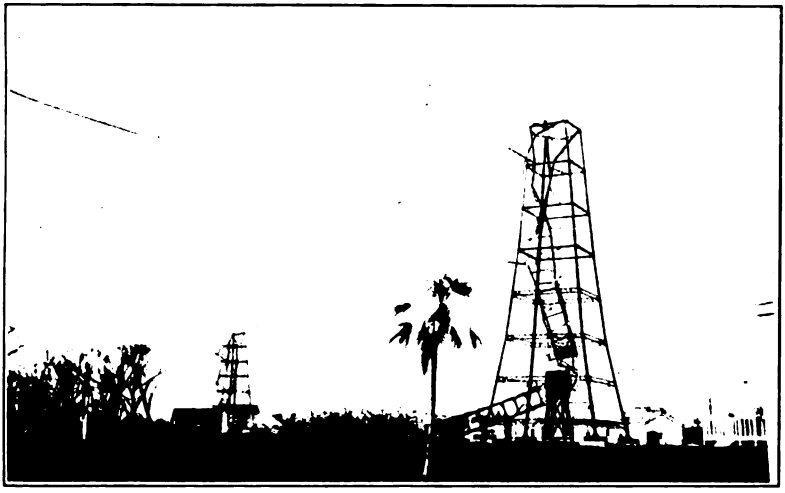

FIGURE 15—View of 1916 wreck, looking north

In a conference held to discuss specifications for the reconstruction of the Swan Island towers, and also for new construction, in the light of the data disclosed by studies of the wrecks, Mr. Davis laid down a new policy, stating, in effect, that:

> "Heretofore it has been more or less customary to take the attitude that when a radio station failed, communication could nevertheless be maintained by means of the cable or wire systems. I want to reverse this idea, and make radio communication so reliable in the countries and districts where we operate that in time of disaster, when all other means of communication fail, the public will be served and can confidently look to our radio stations to maintain service. As long as the radio towers and buildings stand we can give service, and even tho the buildings go, if the towers remain standing, we can quickly fix up a 'jury rig' which will give some degree of communication. It is therefore of the utmost importance that as far as is humanly possible to do so, the towers be designed to withstand hurricanes and earthquakes. Heretofore radio towers seem to have been designed with the principal idea of keeping down the cost, and while we are interested in keeping the cost at a minimum, we are more interested in getting up radio towers which will insure, as far as possible to do so, our ability to maintain radio communication when all other means fail.
>
> "To this end, therefore, we want new towers designed to

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withstand what our experience has taught us are the most severe conditions in the countries and districts where we operate. In short, we want hurricane-proof towers and, from such experience, knowledge and data as we have concerning earthquakes, earthquake-proof towers."

As entirely new material was required for two of the towers above point *G*, and for the other above *E* (Figure 1), (it was neither safe nor economical to use any of the fallen members, under conditions of field work existing at Swan Island), the question of field reinforcement of existing members was confined to about half the leg sections of tower number 4, the cost of which, tho high, seemed justified; and, in accordance with Mr. Davis' instructions, the towers were reconstructed to resist the highest wind velocities. The leg sections above point *G* were strengthened, generally, by the addition of two angles 2½ inches by 2 inches (6.35 by 5.08 cm.), as shown in Figure 16. New horizontal lateral frames were inserted at *P* and *T*-5, and new frames for the messenger cable hitch and masts for receiving wires were added. The guys were renewed on the same plan as has already been described for the reconstruction of two years previous. This work was completed in 1917.

In October, 1921, another hurricane, said to have been one of the most severe ever experienced in that section, swept over Swan Island, but did no damage to the towers and did not even put the station out of commission. This gave us a practical service test, under what may be termed maximum hurricane conditions, and a basis on which we feel considerable confidence in designing towers to withstand storms of the type of West Indian hurricanes. That does not mean that towers so designed would be proof against cyclonic storms, which are of an entirely different character, and for which no reliable data for estimating their force is known to be available. Neither does it mean that provision has been made for resisting other cataclysmic forces of nature, as, with our limited knowledge, such an attempt would either prove futile or very uneconomic. But, during the period from August 15th to October 1st (known as the West Indian hurricane season), in certain of the Caribbean regions destructive hurricanes are of frequent occurrence, and their forces we now seem to be in a fair way to master. It is, of course, possible that we may need yet other lessons, but at least we have made some progress.

Whether the typhoon of the Far East is similar in action and force or not the author cannot state, but he hopes the discussion

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of this paper will throw some light on the subject, as well as that of the force of cyclones.

Some of these stations are located in the earthquake belt of Central America and, naturally, the matter has received some consideration, but no special provision has been made to resist such forces except to take special care to have good foundations and details. Observations on the ground in Guatemala City, in company with engineers who were there at the time of the great earthquakes, impressed the author with the almost insignificant damage done to well-designed steel structures. In fact, most structures well tied together, as with well designed and detailed steel and reinforced concrete frames, escaped with practically no damage to their structural parts.

It is possible that earthquake forces will severely strain the cantilever antenna bridges when they have a considerable overhang, but for these we have used unit stresses only 75 percent of those used for main tower members.

After having been seriously damaged in the hurricanes of 1909 and 1910, Cape San Antonio, Cuba, station was completely wrecked by its third hurricane in August, 1915. This hurricane was unusually severe, but as the wind gauge was carried away early in the storm after having recorded 100 miles per hour (160 km. per hr.), we have no record of its maximum velocity. Mr. John A. (Jack) Cole, the operator in charge of the station, says: "The United States Government barometer was destroyed early in the storm. The Cuban Government barometer, graduated to read from 27.6 to 32.00, was down to its lowest mark; in fact, the indicator was against the pin at 27.6. I do not know how much farther it would have gone if the pin had not been there." Approximate estimates from its effects on the wrecked members indicate a velocity of 120 to 140 miles per hour (192 to 224 km. per hr.)

This station had an umbrella type antenna, with one 250 ft. (76 m.) tower, a duplicate of those at Swan Island, but with four 1-inch (2.54 cm.) steel guys. The station has not been reconstructed. Figures 17 and 18 are views of this tower before and after the wreck. The appearance of the wreck is almost identical with those of Swan Island and helps to confirm the conclusions reached in the analysis of the latter.

Preliminary Design of Three-legged Towers

After the first reconstruction of the Swan Island towers and the final destruction of that at Cape San Antonio, Mr. Davis

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asked for designs and estimates for self-supporting towers, 250, 300, and 350 feet (76, 91.5, and 107 m.) high, with stiff (angle bar) bracing, and alternates for comparison, with adjustable rod bracing. He suggested that three-legged or "triangular" towers be tried for these designs, to which the author replied that, in his opinion, any advantages or saving in weights that might be expected would be more than offset by the increased

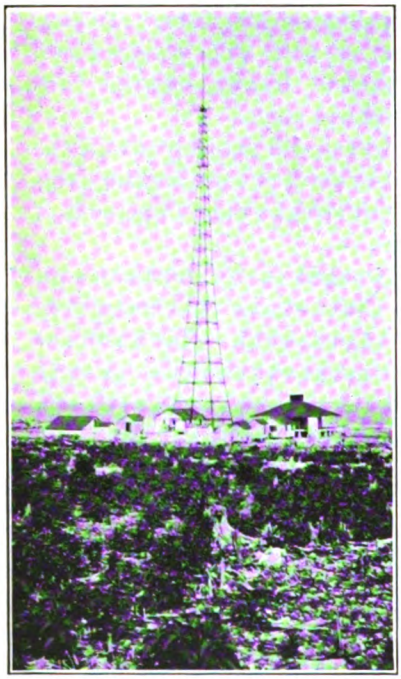

FIGURE 17—United Fruit Company, Cape San Antonio Station before 1915 hurricane

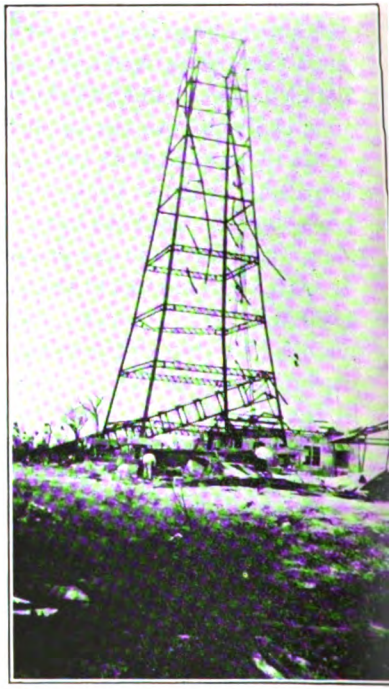

FIGURE 18—United Fruit Compa Cape San Antonio Station in 1915, af hurricane

cost of fabrication, due to the bent plate connections required. Notwithstanding this, Mr. Davis ordered designs for three-legged towers, "to see how they would work out."

The designs were made and the results required the author entirely to reverse his preconceived opinion. There is a saving in weight of about ten percent in favor of the three-legged towe when designed under the same specifications, loads, and uni stresses, as compared with the design for a four-legged tower each tower having its economic width of base. (The side of th

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triangle should be somewhat larger than the side of the square for maximum economy in each case.)

But the three-legged tower has other important advantages: First, the stresses, particularly in the bracing, are not materially disturbed by unequal settlement of the supports, and this is of vital importance where foundation conditions are poor as is frequently the case. It is also important where the tower is carried on distributing girders, which are subject to deflection—a case met with in broadcast stations and others with towers on the top of buildings. Another advantage is the greater rigidity of the triangular cross-section as compared with the square one, thus avoiding all necessity for horizontal "*X*" or cross-bracing.

Computations of areas exposed to wind for a number of cases of comparative designs, showed that stiff, angle-bar, bracing adds about twenty percent to the exposed wind area as compared with a tower with adjustable rod diagonals. Always a large part of the total load, in tropical hurricane sections, where there is no ice load to provide for, the wind on exposed area of tower is the principal force with which to contend, and 20 percent increase in this requires so much additional metal that all advantages of stiff bracing are far outweighed by the increased cost.

Almost all designers of steel structures have used stiff bracing almost exclusively for many years; and for ordinary structures, particularly bridges, it is the correct practice. But the radio tower problem is quite a different case and should be treated on its merits. It is a problem almost unique in the field of framed structures in the fact that the designer has a considerable control over the total of the principal load without reducing the unit loading. No such condition exists to an appreciable extent in bridges or buildings. Moreover, in both bridges and buildings, wind loads are comparatively small and, except in long spans and high, tower-like, buildings, comprise only a very small part of the total stresses.

For these reasons the author has no hesitation in advocating pin connected, adjustable, tension rod diagonals for radio towers. The type has been abandoned so many years for all ordinary steel construction, altho it once was standard even for railroad bridges, that some shops are not well equipped to handle it and have to purchase the rods outside. Moreover, it is a slightly more expensive class of shop work and therefore costs more per pound—but there are not so many pounds. Not only is the bracing itself lighter but the legs are also very much lighter because the loaded wind area is much less, and consequently the leg

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stresses are materially less. Total cost and not cost per unit is what spells economy. In one case we were able to pay a very much higher price per pound for the steel and still save nearly 20 percent on the cost of the completed towers, due to saving in weight of steel and of one foundation and pedestal per tower.

Another feature introduced in all later designs is that the leg sections are all equal short chords of a perfect circular arc. A little pains taken in fixing the outline, so that leg sections are all of equal lengths, permits the fabrication of these members as duplicates and interchangeable thruout the height of the tower, except where sectional areas have to be adjusted to the stresses, and, of course, excepting the top and bottom sections. This feature makes the fabricating cost of these legs as low as any straight work. That the vertical height of the stories is slightly variable is of no consequence whatever. It is true that the angle between the leg and horizontal strut is different at each full story, requiring special lay-out and template for each point, but as an objection to the design this loses its force when considered in connection with the fact that each tower has at least six duplicate bent plates of each kind; and as several towers are generally fabricated together, on the same order, the duplication is ample to distribute the cost of drawings, special lay-outs, and templates over a sufficient number of duplicate pieces to keep the unit cost down to a low figure. There are now under order eight duplicate towers, for example, which gives forty-eight duplicate bent plates from each template, so that the cost of the items mentioned for each piece cannot be very high. Moreover the angle of the bend is only 15 degrees and, for all practical purposes the slight variation from that figure from bottom to top is negligible, so that all can be made with the same bend angle, and this is the only expense chargeable to the three-legged feature.

Recent Designs

All the features herein described were used in the design for the towers of the Almirante, Panama, station, built in 1921, and illustrated in Figures 19 and 20, showing, respectively, stresses, sections, general data, and typical details. Figure 21 is a view of the completed towers.

These towers have no antenna bridges, spreaders being used instead.

Most fabricators, at first sight, are inclined to exaggerate the difficulties and cost of these designs, but those who have been

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properly equipped have found out after fabricating a few that the shop costs were not as high as they anticipated, so that now there is no trouble in obtaining satisfactory prices. The unit prices secured for the later towers are only slightly higher than is asked for towers on fabricators' designs, leaving a substantial net saving in total cost due to the more economical use of steel.

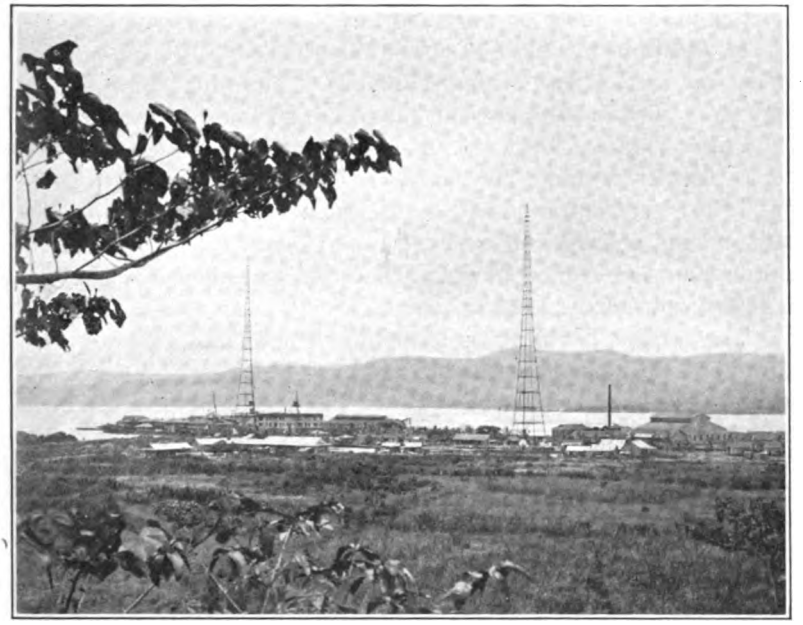

FIGURE 21—United Fruit Company, Almirante Station, 1921

There are, however, a few fabricators who sky-rocket their quotations at the sight of a curved outline or adjustable tension rod, and these are the ones most heard. The type of design preferred by them has four legs, either straight or with only one break in the height of the tower, and stiff bracing; extremely slender struts and, generally, metal only 3/16 inch (0.48 cm.) thick—sometimes only ⅛ inch (0.32 cm.) thick—except for the legs.

Such designs have been perfectly described by the Dean of American Bridge Engineers, Mr. Gustave Lindenthal, in a paper entitled "Some Thoughts on Long Span Bridge Design," (1921), in these words:

> "* * * * * an American design, a typical shop creation, * * *" and "Ugliness is rational beauty to minds who see merit only in geometrical figures and in cheap fabrication."

and, it is to be noted that he was discussing the design of steel towers.

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Antenna Bridges

When, in place of spreaders and messenger cables, fixed antenna bridges came to be required on top of towers, we followed the same general idea:—that of keeping the area exposed to wind down to a minimum. The importance of this increases as the height of the area under consideration and, consequently, its overturning lever arm, increases. It is further intensified where hurricane force must be provided for.

In addition to the use of adjustable rods for all tension members, we used extra strong, galvanized pipe for compression members, in two symmetrical, balanced, suspension cantilevers. The area exposed to wind, including about half that of the leeward truss, is under 200 sq. ft. (18.6 sq. m.) for the 150 ft. (46 m.) of bridge. Compared with fabricators' structural steel designs this means a saving of over sixty percent in the wind load on antenna bridge—quite an item when one considers it acting on a lever arm 400 ft. (122 m.) high.

The trusses act only as stiffening trusses and to brace the lower chord which takes the outboard thrust from the links. The walkway floor is of Irving Subway Grating, type "*L*," ¾ inch (1.9 cm.), which was adopted because it combines with minimum weight, ample resistance to lateral shear, avoiding the use of lower laterals, and with the maximum size of mesh. Tests made by the Irving Iron Works Company, Long Island City, especially for this work, showed that, at the elastic limit, the shearing strength of this grating is 393.7 lbs. (179 kg.) per mesh width of 2⅜ inches (6.03 cm.); and the ultimate strength 448.1 lbs. (220 kg.) per mesh; the tests being made on panels 26⅛ inches (65.2 cm.) wide, having 12 straight bars and 11 meshes of 2⅜ inches (5.95 cm.). Figure 22 shows panel number 2 after testing.

The first towers of the Tropical Radio Telegraph Company with fixed antenna bridges are those erected in 1922 at Tegucigalpa, Honduras. Eight similar towers, for four stations, are now under order.

Figure 23 shows the stresses, sections, and typical details of the Tegucigalpa towers, Figure 24, the antenna bridge, and Figure 25 a view of one of the towers after completion. Figure 26 is a view looking thru the antenna bridge of the East tower at Tegucigalpa.

Figures 27 and 28 are, respectively, the stress, section, general data sheet, and general plan and details, with stresses and sections of the antenna bridge, for the latest, revised standard 400 ft. (122 m.) self-supporting tower with 150 ft. (46 m.)

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antenna bridge. A comparison of Figures 23 and 24 with 27 and 28 will give some idea of the advance made in one year in the tower practice of these companies. The principal difference is the antenna system provided for, as indicated by the plans, but what may not be clear without explanation is that most of the other changes were required by the increase in wire pull and provision that the pull may be in either direction. Some simplification of details of bridge trusses, particularly in the center panels at the tower, are about the only other changes.

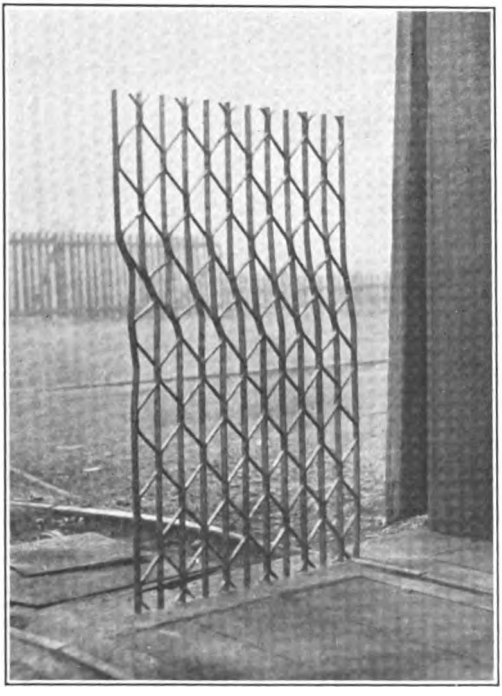

FIGURE 22—Irving Subway Grating

WIND PRESSURE

The wind pressure used for these designs was adopted after a study of all accessible data, but, as our records of the Swan Island and Cape San Antonio hurricanes seem best and most reliable for the conditions required in those localities, they were naturally given the most weight in specifying wind loads. As has been stated, these records indicate probable velocities of about 137 miles per hour (220 km. per hr.), or a maximum pressure on small areas of 75 lbs. per sq. ft. (368 kg. per sq. m.), derived by

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formulas from the computed breaking load of 50 lbs. per sq. ft. (245 kg. per sq. m.), average, on the 100 feet (30.5 m.) of tower that broke off, and taken on 1½ times the area of the projected elevation of one side. The conversion formula used is $P = 0.004V^2$ in which P = pressure in pounds per square foot, and V = velocity in miles per hour. The United States Weather Bureau formula is $P = 0.004V^2$. $B/29$, in which B = barometer reading in inches.

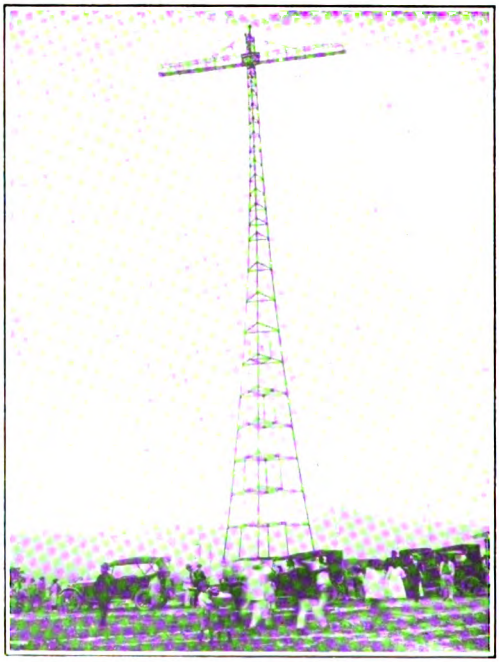

FIGURE 25—East Tower, Tegucigalpa (Honduras), Radio Telegraph Station of Tropical Radio Telegraph Company, erected 1922

Some authorities give $P = 0.0042\ V^2$, and an older formula is $P = 0.003\ V^2$, but the latter is probably nearer the average pressure on large areas than the maximum on small areas which is desired. A number of more complicated formulas have been proposed, some based on rational theories, but they have yet to be proven more correct.

Many authorities make large allowances for reduction in velocity and pressure as ground level is approached, compared with those at high elevations above ground; but while many observations, particularly those on the Eiffel Tower, have shown this to be correct in light winds, they have shown also that at higher

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velocities, "the pressure at the top and that at the ground come nearer together." At the highest velocities it is probable that there is not much difference, and it is only the highest velocities that interest the designer. Until it is shown that high winds and hurricanes have similar properties and effects as light winds, which has yet to be done, we should not give too much weight to deductions from observations in light winds. Unfortunately very few records are available for very high winds, 96 miles per hour (154 km. per hr.), being the highest reported in observations covering six years at the Forth Bridge—equivalent to barely 37 lbs. per sq. ft. (181 kg. per sq. m.). The highest velocity reported by Mr. S. P. Wing in "The Electrician" (London), of July 1st, 1921, was 37 miles per hour (59 km. per hr.), equivalent to less than 5.5 lbs. per sq. ft. (27 kg. per sq. m.). These cannot safely be taken to indicate the characteristics of winds from 100 to 140 miles per hour (160 to 224 km. per hr.), or 40 to 80 lbs. per sq. ft. (196 to 392 kg. per sq. m.).

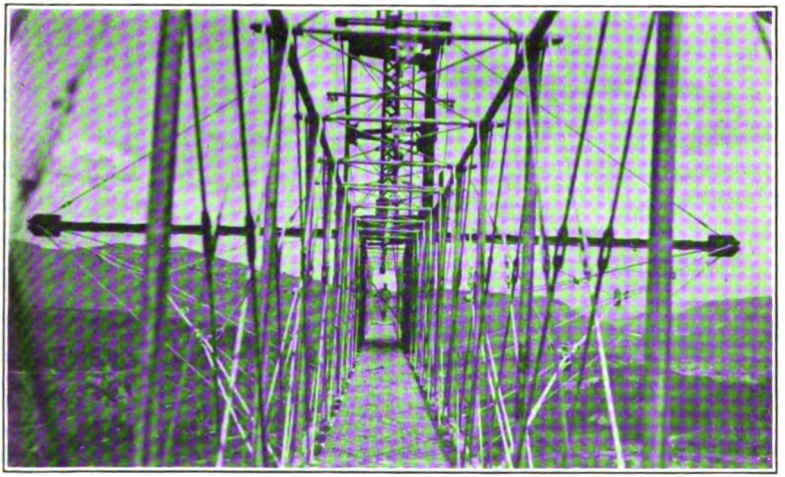

FIGURE 26—Looking through 150 ft. (46 m.) Antenna Bridge, Tegucigalpa

Sir Benjamin Baker, in reporting the Forth Bridge observations, says, "It was found that the large gauge (15 feet by 20 feet (4.6 by 6.1 m.) invariably recorded smaller average pressures than the small, circular, gauges (1.5 sq. ft.) (0.139 sq. m.), never exceeding 80 percent of the latter for observations at the same time, and generally the average pressure on the large gauge varied from 50 to 70 percent of the pressure indicated on the small gauges. Furthermore, small gauges installed in

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[illegible] of the [illegible] tower of the bridge showed [illegible] twelve pounds and to 5.47 kg. in pressure, [illegible] and sometimes the other showing the greater [illegible]. The large area does not show more than two-[illegible] of the pressure indicated by the small gauges as an [illegible].

[illegible] on Wind Stresses" and an article in "En-[illegible]" of March 1[illegible] both by Mr. Robins [illegible] some interesting data but nothing on hurri-[illegible] and not much on very high winds—nothing [illegible] to the case as the Forth Bridge obser-vations.

The [illegible] now specified for towers in the hurricane belt—Gulf and Caribbean sections—are as follows:

80 lbs. per sq. ft. 3.42 kg. per sq. m. on surfaces— used to fix the scale.

70 lbs. per sq. ft. 372 kg. per sq. m. — on the elevation of bridge on a plane 87° from center line of bridge).

68 lbs. per sq. ft. 333 kg. per sq. m. — on $1^1/_2$ times the projected side elevation of bridge.

65 lbs. per sq. ft. 333 kg. per sq. m. — on $1^1/_2$ the projected elevation of top story of tower.

and 150 lbs. per vertical ft. 226 per vertical m. of height of tower to lower chord of bridge.

The last item works out as follows: at $1^1/_2$ times projected elevation.

61.0 lbs. per sq. ft. (295 kg. per sq. m.) on upper five (5) stories.

56.8 lbs. per sq. ft. (278 kg. per sq. m. on upper nine (9) stories.

50.5 lbs. per sq. ft. (248 kg. per sq. m. on upper thirteen (13) stories.

43.2 lbs. per sq. ft. (212 kg. per sq. m.) the entire tower from base to lower chord of bridge).

These reductions were arrived at empirically, as no rational formula available has been proven correct. They are intended to provide both for the reduction in average intensity on large areas, and also for some reduction of velocity as ground elevation is approached, but the influence of the latter is considered of only secondary importance.

The specified "150 lbs. per vertical foot (226 kg. per vertical [illegible]) of height" was only adopted after a number of computations loadings per sq. ft., progressively decreasing from top to base,

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two opposite corners of the central tower of the bridge showed variations of ten to twelve pounds (4.6 to 5.45 kg.) in pressure, sometimes the one and sometimes the other showing the greater registration. The large area does not show more than two-thirds of the pressure indicated by the small gauges as an average."

"Six Monographs on Wind Stresses" and an article in "Engineering News-Record" of March 16, 1922, both by Mr. Robins Fleming, contain some interesting data, but nothing on hurricane velocities and not much on very high winds—nothing appearing so applicable to the case as the Forth Bridge observations.

The wind loads now specified for towers in the hurricane belt—Gulf and Caribbean sections—are as follows:

80 lbs. per sq. ft. (392 kg. per sq. m.) on antenna—(used to fix the sag).

76 lbs. per sq. ft. (372 kg. per sq. m.)—(on the elevation of bridge on a plane 87° from center line of bridge).

68 lbs. per sq. ft. (333 kg. per sq. m.)—(on 1½ times the projected side elevation of bridge).

68 lbs. per sq. ft. (333 kg. per sq. m.)—(on 1½×the projected elevation of top story of tower).

and 150 lbs. per vertical ft. (226 per vertical m.) of height of tower to lower chord of bridge.

The last item works out as follows: (at 1½ times projected elevation).

61.0 lbs. per sq. ft. (295 kg. per sq. m.) on upper five (5) stories.

56.8 lbs. per sq. ft. (278 kg. per sq. m.) on upper nine (9) stories.

50.5 lbs. per sq. ft. (248 kg. per sq. m.) on upper thirteen (13) stories.

43.2 lbs. per sq. ft. (212 kg. per sq. m.) (the entire tower from base to lower chord of bridge).

These reductions were arrived at empirically, as no rational formula available has been proven correct. They are intended to provide both for the reduction in average intensity on large areas, and also for some reduction of velocity as ground elevation is approached, but the influence of the latter is considered of only secondary importance.

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These reductions were arrived at empirically, as no rational formula available has been proven correct. They are intended to provide both for the reduction in average intensity on large areas, and also for some reduction of velocity as ground elevation is approached, but the influence of the latter is considered of only secondary importance.

The specified "150 lbs. per vertical foot (226 kg. per vertical m.) of height" was only adopted after a number of computations of loadings per sq. ft., progressively decreasing from top to base,

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had shown that a uniform load per ft. of height is as near to the probable actual load as any that can be devised with our present limited knowledge, and has the advantage of simplicity with many precedents in bridge specifications. For stiff bracing 180 lbs. per vertical foot (249 kg. per vertical m.) was found to give the equivalent loading per square foot for the Almirante towers.

As has been stated, the Swan Island towers and that at Cape San Antonio failed at about 100 ft. (30.5 m.) from the top, under a wind load very close to 50 lbs. per sq. ft. (245 kg. per sq. m.), as an average, uniformly distributed over 1½ times the area of the projected elevation of the section involved. Also that practically no damage was done to the lower 75 ft. (22.9 m.) of the towers; and that between 75 ft. (22.9 m.) and 150 ft. (45.8 m.) in height the destruction was caused by the stripping action of the falling tops. If these deductions are correct—and we have no better data—the loading adopted is safe and has been so proven to a considerable extent by the hurricane of 1921. Altho the specified loading is only equivalent to 43.2 lbs. per sq. ft. (212 kg. per sq. m.) on the area involved for the lower story members, it is considered safe since these members are good for 60 to 65 lbs. per sq. ft. (284 to 319 kg. per sq. m.) within their elastic limit. This area is over 1,500 sq. ft. (139 sq. m.) (extending 400 ft. (122 m.) vertically and 150 ft. (45.7 m.) laterally), nearly seven times that involved at the initial failure at Swan Island and five times that of the larger gauge at Forth Bridge, with outside dimensions ten to twenty times the latter. The maximum wind pressure assumed is double that of the highest recorded by Sir Benjamin Baker, who says, "The higher wind pressures come more in gusts than in even pressure extending over a large area," which should be given great weight until more comprehensive observations supersede it. With a double pressure it is reasonable to assume the wind more gusty, and the average pressure a smaller percent of the maximum. Also, with the much larger area and dimensions, this factor should be less. Therefore, instead of the average pressure on an entire 400 ft. (122 m.) tower being 2/3 of the maximum, it should be, in all probability, very much less. While we have designed the lower stories for an average pressure only 54 percent of the maximum, within the elastic limit they are good for about ¾ of the maximum, which is considered as large a factor of safety as is warranted, in view of the cost.

In sections where the highest recorded winds do not exceed 90 to 100 miles per hour (144 to 160 km. per hr.), loads of 40 lbs.

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per sq, ft. (196 kg. per sq. m.) at the top with small areas involved, graduated down to 25 (123), or, for towers 600 ft. (183 m.) high and over, to 20 lbs. per sq. ft. (98 kg. per sq. m.) for the lower members will meet the average case. In many of these cases an ice load must also be provided for.

Where municipal building codes have to be complied with, some of which require absurdly low unit stresses for wind loads, something may be gained by absorbing the ice load in the excessive "safety factor."

Wire Pull, Torsion, and Direction of Pull

For wire pull on towers, wind and ice loads on antennas have been provided for by using the strength of the wires, these only being used to compute the sag required for the wires to take the loads. Where high unit stresses are used for the tower members the breaking strength of the antennas may be used, but where towers are designed with low unit stresses, as required by some building codes, the elastic limit of the antennas may be safely used for the maximum wire pull on tower, as the low unit stresses provide an ample margin for the excess load due to wires being pulled so taut as to develop their ultimate strength; but with high unit stresses, 24,000 lbs. per sq. in. (3.720 kg. per sq. cm.), caution should be used not to impair the margin of safety.

Some provision for future increase of wire pull may be advisable, as in the latest standard design of the Tropical Radio Telegraph Company, where an increase of 50 percent is anticipated. The wire pull used for tower design increased from 2,000 lbs. (920 kg.) in 1906 to 14,000 lbs. (6,350 kg.) in 1922, and 32,000 lbs. (14,500 kg.) has been used for high power, intercontinental stations. In one case a change in antenna plan increased the wire pull 30 percent before the towers were erected—an indication of the necessity for some margin in many cases. About the least margin should be to use the ultimate strength of the largest number of heaviest wires proposed at the time, which might permit 20 percent increase, assuming elastic limit at 2/3 the ultimate strength.

For torsion it is customary to consider an unbalanced load equal to the three wires nearest the end of antenna bridge as out. It is extremely improbable that more than three wires would go out on one side before any on the other side went out. To provide for an unbalanced load of all the wires on one side would require a wider top and add considerable to the cost. In the standard design, Figures 27 and 28, the top width has been in-

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creased to 10 ft. 6 in. (3.2 m.) from 7 ft. 0 in. (2.1 m.) of the Tegucigalpa towers, Figures 23 and 24, principally on account of the greater torsion due to increase in size and arrangement of the wires. However, excessive torsional stresses will be relieved to some extent by the elastic twisting of the towers, so that they could not equal the stresses as computed by rigid statics. The elastic equilibrium between the twist of the towers and the sag, stretch, and pull of the antenna can be computed, but it is a tedious problem, and a small variation in the wires, their arrangement and initial sag, would entirely change the result and the computations would then be valueless.

Several devices have been designed to release an entire line of wires when a break occurs in any one span, which, were it safe to count on it always being maintained in working order, would permit a considerable saving in weight and cost of towers; but that seems a rather unsafe assumption, as it is so easy and simple for any one to make a wire fast at any time, and would require constant vigilance to make sure it was never done. The only safe assumption is to consider the antenna as fixed at each point of support.

With the ends of the antenna held by counterweights or supported from triatics (instead of antenna bridges), the maximum wire pull could be very greatly reduced. Such devices may be made reasonably secure against tampering, but the tower designer must take the antenna plan from the radio engineers, just as the bridge designer must use the rolling loads fixed by the locomotive and car builders.

Under maximum wind load the catenary of the antenna lies in a nearly horizontal plane, and the tangent to the curve at the point of support is the direction of the pull on the tower. For this reason, in a two-tower station, the apexes of the triangles should face each other, as this gives the most favorable distribution of loads on the legs of a three-legged tower. Altho some experienced engineers set the towers with the sides of the triangular cross sections facing each other, instead of the apexes, precise analyses show that such an arrangement increases the maximum compressive stress due to wire pull in the legs, in some cases as much as 20 percent.

With four-legged towers, however, the sides and not the corners should face each other for most economical distribution of compressive stresses in the legs, the difference being more pronounced than in three-legged towers and, therefore, a more important matter. In other words the stresses in three-legged

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towers are less sensitive to direction of wire pull, because, for one reason, even with the most favorable arrangement the maximum leg stress is a larger proportion of the total and, consequently, subject to less variation due to change of direction of wire pull; which is another point in favor of three-legged towers.

Stresses, Sections, and Details

The total wire pull and its direction having been determined, and also the resultant if the tower carries more than one span or set of antenna wires, a convenient method of stress analysis is to find the resultants of wire pull, wind loads on top story of tower, and wind loads on the antenna bridge if there is one, and then find the components of these resultants normal and parallel to the center line joining the towers, as shown in Figure 27. This must be done for several directions of wind, as no one direction will give the maximum for all members. The maximum wire pull due to wind occurs with wind normal to the axis of the antenna, but this will rarely give the maximum combination of simultaneous stresses in a member. For other wind directions the wire pull may be reduced by the co-efficients of direction.

In combination with torsion the total wire pull should be reduced in the proportion that the number of wires remaining in bears to the full number. The principal effect of torsion is on the diagonal rods in the upper stories where the lever arm of resistance is small. In the legs torsional stresses are not cumulative, but are balanced in each story.

In computing stresses, some prefer one method and some another. Those accustomed to analytical methods and the slide rule may find, as did the author, that it pays to compute an influence table for all members, or the stress co-efficients for unit loads, both at line of resultant of the top loads and for the center of action of wind on the part of tower involved for each member. If the method of moments is used, the lever arms may be found either graphically or computed from similar triangles by slide rule. It is necessary to find the stresses in the legs for at least two conditions to get the maximums of compression and tension. The maximum combinations of stresses from the several loads, which must be found, are somewhat complicated, so that, until one has computed a few similar towers, it may be necessary to try several cases to be sure of having the maximum combination.

The unit stresses used to fix the sections of tower members are specified in Figures 23 and 27, and those for the antenna bridge members in Figure 28. The latter are lower than those

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for the tower members for the following reasons: (1st) The bridge is subject to more sudden loading, being the first part of the structure to receive the wire load, in which there may be an element of impact. (2nd) It will be subject to heavy loading more frequently and is more liable to maximum loading than the tower members as its total loads are composed of fewer elements, while in tower members proper a full maximum load can only occur when three or four elements or conditions occur simultaneously with maximum wind—a rare probability. (3rd), and last, but not least, an excess wire pull will cause much larger proportionate excesses in stresses of bridge members than in tower members.

The unit stresses for tower members are justified by precedent, by many authorities and by the merits of the case: A base unit stress of 24,000 lbs. per sq. in. (1,710 kg. per sq. cm.) has been used for many years for the lateral wind bracing of highway bridges. It is permitted for wind loads by the building codes of several cities, including New York. It is now proposed for wind stresses in members carrying no load stresses (dead or live) for railroad bridges by the board of engineers of the American Institute of Steel Construction. Practically all stresses in both tower and bridge are due to wind and dead loads, the dead load stresses being only a small part of the total, particularly for the tower members.

The sections for the legs are built up of three angles, except those of the two top stories, and are economical and compact, with not too large area exposed to wind. This type of section is good for towers up to about 450 ft. (137 m.) in height designed for hurricane force, and for towers up to about 600 ft. (183 m.) designed for the ordinary wind loading of 30 lbs. per sq. ft. (147 kg. per sq. m.), average. Higher towers require sections of different design.

The sections of most all the horizontal struts are fixed by the ratio of slenderness (length divided by radius of gyration), those in the lower stories requiring sub-struts and hangers to reduce the unsupported length. In towers much exceeding 400 ft. (122 m.) in height, the economic width of base is too much for the economic use of horizontal struts, and for export work the members would be so long that shipping difficulties would be encountered. In such cases resort has been had to some form of "A" frame bracing, similar to that shown in Figure 32. Possibly a better, and certainly a more graceful solution of this problem would be a base section in an arch form, for which the

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gothic-lancet lines seem most adaptable. Such a solution, however, would require very painstaking work in detailing, to avoid fabricating difficulties and high cost.

Figure 33 shows a plan proposed but never executed, for constructing reinforced concrete tower bases arranged as station buildings. The plan is not economical and is only available for locations where an architectural effect, with elaborate station buildings, are desired.

All details, even the minor ones, should receive the same degree of care and attention as the most important member, or satisfactory results and even safety will not be attained. The details of the anchorages, and the pin connections of the diagonals must be efficient and mechanically correct; splices must be centric and proportioned for the full strength of the members connected; and every detail must be planned with a view to facilitating the transportation and erection.

The diagonals have right and left threads on upset ends, with clevis nuts on cotter pins. Socket type safety set-screws in the clevis nuts serve to lock the rods after adjustment, as a precaution against the tendency of all maintenance men to screw up all slack rods, ignorant of the fact that some of the rods should be slack when there is any wind blowing, and to pull them up simply throws the tower out of line when the wind changes. It is a very difficult job, requiring an expert, to line up a crooked tower under wind—and the wind blows a stiff breeze most of the year in the trade wind belt.

Broadcast Station Towers on Buildings

Two pairs of 100 ft. (30.5 m.) towers for broadcast stations, designed on the lines of the Fruit Company's standard plan, have recently been constructed for the Radio Corporation of America, one pair in Washington, D. C. (station WRC), and one on Aeolian Hall, New York (stations WJY-WJZ), the latter shown with permission in Figure 29. The only feature of these broadcast station towers which departs from the lines already described is the base girder system, introduced to distribute the loads to the structural frame of the building. In the case of Aeolian Hall, an existing building, it was necessary to distribute the load on a number of columns to avoid reinforcing the latter down thru the building at a prohibitive cost. At first glance it appeared that four-legged, square base, towers would fit best on a building with rectangular bays, but the three-legged towers with triangular bases actually proved

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gothic-lancet lines seem most adaptable. Such a solution, however, would require very painstaking work in detailing, to avoid fabricating difficulties and high cost.

Figure 33 shows a plan proposed but never executed, for constructing reinforced concrete tower bases arranged as station buildings. The plan is not economical and is only available for locations where an architectural effect, with elaborate station buildings, are desired.

All details, even the minor ones, should receive the same degree of care and attention as the most important member, or satisfactory results and even safety will not be attained. The details of the anchorages, and the pin connections of the diagonals must be efficient and mechanically correct; splices must be centric and proportioned for the full strength of the members connected; and every detail must be planned with a view to facilitating the transportation and erection.

The diagonals have right and left threads on upset ends, with clevis nuts on cotter pins. Socket type safety set-screws in the clevis nuts serve to lock the rods after adjustment, as a precaution against the tendency of all maintenance men to screw up all slack rods, ignorant of the fact that some of the rods should be slack when there is any wind blowing, and to pull them up simply throws the tower out of line when the wind changes. It is a very difficult job, requiring an expert, to line up a crooked tower under wind—and the wind blows a stiff breeze most of the year in the trade wind belt.

Broadcast Station Towers on Buildings

Two pairs of 100 ft. (30.5 m.) towers for broadcast stations, designed on the lines of the Fruit Company's standard plan, have recently been constructed for the Radio Corporation of America, one pair in Washington, D. C. (station WRC), and one on Aeolian Hall, New York (stations WJY-WJZ), the latter shown with permission in Figure 29. The only feature of these broadcast station towers which departs from the lines already described is the base girder system, introduced to distribute the loads to the structural frame of the building. In the case of Aeolian Hall, an existing building, it was necessary to distribute the load on a number of columns to avoid reinforcing the latter down thru the building at a prohibitive cost. At first glance it appeared that four-legged, square base, towers would fit best on a building with rectangular bays, but the three-legged towers with triangular bases actually proved

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more adaptable. Each tower is supported on ten building columns, so that the load added to any one column was within the limits allowed by the code and regulations of the Building Department. Figure 30 is a view of the towers on Aeolian Hall.

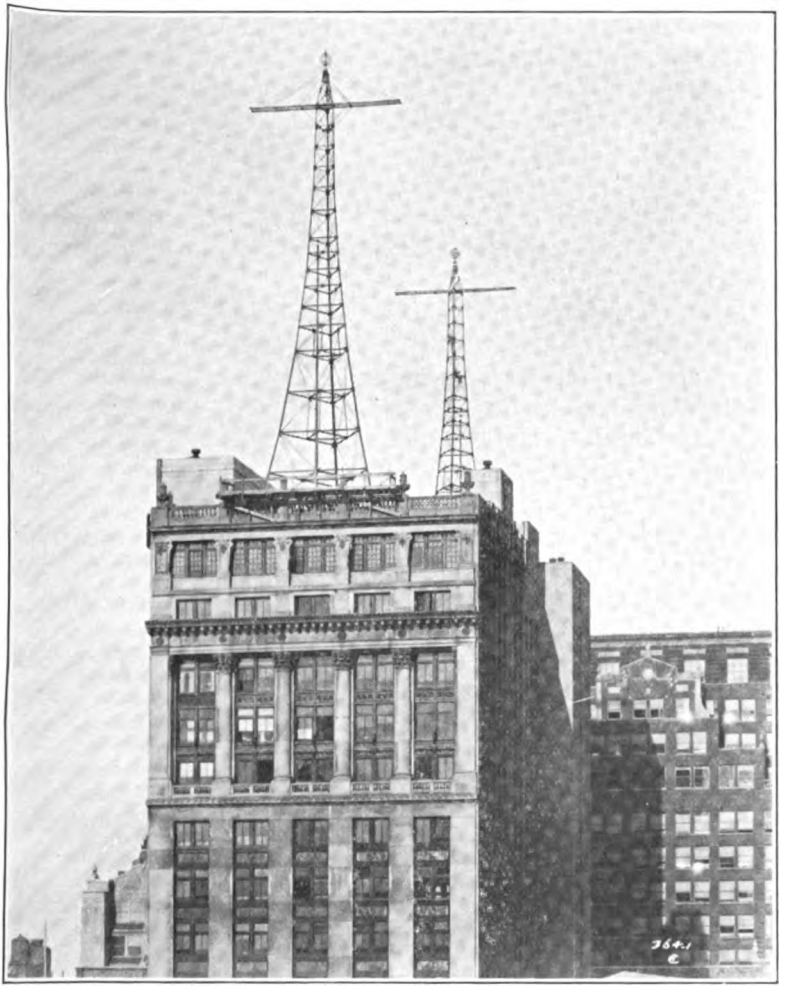

FIGURE 30—Broadcast Central, WJY-WJZ, Radio Corporation of America, at Aeolian Hall, New York

The Washington Station presented quite different conditions. There the buildings were of reinforced concrete under construction and the columns were designed to carry four additional future stories—giving ample column capacity for the tower

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loads. Some complication was met with in the irregular dimensions of the bays and their skew of about 30 degrees due to the street angles. Here, also, the three-legged towers proved the more adaptable, but required a base width of 24 ft. (7.3 m.) instead of the 28 ft. (8.5 m.) base of the Aeolian Hall,NewYork, towers, where the building bays are 14 ft. 2 in. (4.3 m.) wide.

One of the Washington towers is carried by four building columns and the other by five, but not equally distributed as to loads. Three columns would have been ample for the loads of one tower, but their arrangement did not permit a symmetrical base. This was met by putting in distributing girders, resting on the building columns, and supporting the towers; the three points of support on the girders forming equilateral triangles.

While one pair of towers required a base width of 28 ft. (8.5 m.) and the other pair 24 ft. (7.3 m.), the leg sections are duplicates for all and were made from the same drawings and templates, which are also designed to serve for other 100 ft. (30.5 m.) towers with bases of 20 ft. (6.1 m.) and 16 ft. (4.9 m.) should they be required for other locations. Sixteen feet (4.9 m.) is about the economic width of base for three-legged towers 100 ft. (30.5 m.) high, considering the tower alone, but when it is necessary to fit the bays of a building and keep the loads on the columns down, it will generally be found cheaper to get the spread in the legs of the tower rather than to make the base or distributing girders excessively heavy.

For the tower itself, the economic width of base is that which makes the weight of the legs about equal to the weight of the bracing.

The antenna bridges of these broadcasts station towers are 36 ft. (11 m.) long and 3 ft. (0.92 m.) wide, formed of two lines of 3½ inch (8.87 cm.) extra strong, galvanized pipe, connected together by Irving Subway Grating, and supported by truss rods to the top of the tower. The grating not only serves as a walkway, but also to take the shearing stresses due to wire pull and wind. The bridges on the Washington towers are set perpendicular to a side of the towers, making the antenna axis parallel to a side, which arrangement is not economic as it increases the stresses and sections of the front legs so that they have to be made specials and not duplicates of the other legs. It also increases the total weights somewhat, but it did not seem to be avoidable in this case on account of the skew angle of the buildings.

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Semi-guyed Towers and Guyed Steel Masts

The Swan Island towers as reconstructed and shown in Figure 5 are, in fact, semi-guyed or semi-self-supporting, but this plan was only adopted as an economic expedient for salvaging a considerable part of the old steel which had not been injured.

The question has been asked—are there any advantages in a semi-guyed tower for new construction? The answer, as a general proposition, is no; but there may be rare conditions where the relative costs of the several kinds of material and facilities for putting them in the work are exceptional, which might, possibly, give some advantage to such a design. The author, however, has yet to meet such a case. The first cost of the guys, if put into additional material in the legs, would make the towers nearly if not quite self-supporting and, moreover, the maintenance and renewal of guys is many times more costly than of structural steel members. Guys increase the number of parts and also require anchorages, and they generally have to be insulated. They occupy a larger ground area, and probably increase the absorption of power.

Economically the problem has no intermediate solution:—the only alternative to the self-supporting tower that has any advantage at all is the guyed mast, and the only advantages of the guyed mast are rapidity of erection and low first cost.

The United Fruit and Tropical Radio Telegraph Companies have a number of guyed steel masts in service. Figure 31 is a view of the New Orleans station, showing four guyed steel masts. Maintenance records show that, in the tropics, guys require renewal about every five years, and that after the second renewal the cost is more than the self-supporting towers. It has been stated that, in dry climates, a set of guys is good for eight years. If these estimates are correct, the self-supporting tower is the cheaper after ten years in the tropics and after sixteen years in dry climates.

Disadvantages of guyed masts, in addition to the greater cost of maintenance, are: greater area of ground required; absorption of power by the guys, even when insulated; and impossibility of attaching an antenna bridge. Often, therefore, two guyed masts are required to do do the work of one tower.

Design for 1,000 Foot (305 m.) Tower

Figure 32 shows a design for a tower 1,000 feet (305 m.) high, which, however, was not built. It was proposed for three-tower stations, with the antennas supported on messenger cables and

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equalizing bridles in place of spreaders, with a maximum wire pull of 40,000 lbs. (18,400 kg.). With this layout, there would always be compression in the front leg and tension in the rear legs except when the mat might be blown out, in which case there would be no wire pull. This condition of stress distribution, requiring especially heavy sections for the front leg—the main compression member—suggested the attempt to introduce in a self-supporting tower the mast element, substituting stiff legs for guys. The front leg, as shown in the figure, is a straight, vertical line—the most favorable for a heavy compression member, other things being equal (which generally they are not), and the rear stiff legs are subject to tension only when under full load, and therefore require only sufficient compressive strength to hold the tower when the mat is out. The rear, stiff legs are curved to reduce the weight of the bracing, and to add flexibility to the tower.

FIGURE 31—STATION OF THE TROPICAL RADIO TELEGRAPH COMPANY AT NEW ORLEANS, ERECTED 1914

The plan worked out well on paper, the weight estimate showing it to be very economical in material, but it will probably never be used in just this form, since the antenna lay-out which led to it has been superseded, and also because no two members in such a tower could be made duplicates except the rear leg sections and some of the diagonals, and of these there would only

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be two pieces alike in a tower, thus greatly increasing cost of fabrication.

The photographs of the Swan Island wreck of 1916 were taken by Mr. H. O. Easton, Superintendent, Radio Department, United Fruit Company, whose comprehensive selection of views, with his report on the wreck, made possible the studies of wind pressure in hurricanes.

Lieut.-Col. W. P. Rothrock, Member of the American Society of Civil Engineers, assisted in the analysis of the 1916 wreck, superintended the erection of the Almirante, Panama, and the Tegucigalpa towers, and suggested a number of improvements in details which make for efficiency and facilitate erection.

The fabricators and their engineers have given fine co-operation in solving problems of details and construction. Included in the list are Baltimore Bridge Works (now Carnegie Steel Company, Baltimore Warehouse); Virginia Bridge and Iron Company; Fort Pitt Bridge Works; Dietrich Brothers, Baltimore, Maryland, and the McClintic-Marshall Company (Riter-Conley Manufacturing Company).

Conclusion

In conclusion it may be stated that nothing has been added to any of the towers described in this paper for the purpose of producing effects or pleasing appearance, except the skeleton globes on top of the 100 ft. (30.5 m.) broadcast station towers of the Radio Corporation of America, which globes are practically that corporation's trade mark. The observation platform and railing at the 150 ft. (46 m.) level on the design for the 1,000 ft. (305 m.) tower has a self-evident purpose. Neither has line or form been influenced by any consideration except the requirements of the mathematical statics and economics of the problems, save for the scroll railing braces at the ends of the antenna bridges.

In the preliminary design of the 150 ft. (46 m.) antenna bridge, the ends were shown cut off square at the panel point, the blunt appearance of which drew a criticism from Mr. Davis, who asked if something could not be added to improve it. The author's reply was that he had tried several ways to improve this, each of which only made it worse, and that nothing could be added which lacked an obvious purpose without detriment to appearance. Up to that time, the railing braces seemed to serve no useful purpose, but when worked out in detail it was evident that a man fixing the end antenna would be in a difficult and

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dangerous position without these braces. Therefore the lower chords were extended 3 ft 9 in. (1.15 m.) beyond the panel point to support a two foot (0.65 m.) extension of the walkway and braces to a railing around the end. The scrolls in the railing braces are a conventional and common treatment of this detail. In Figure 26 a man may be seen sitting on the end railing.

The appearance of these towers has been commended by many people of taste, including some architects. This is greatly appreciated but, if it is merited, it should be credited entirely to the effect of the theorem that a structure designed by the rules of statics and economics, if the solution is truthful, will conform to all the rules of aesthetics, and vice versa. It is hoped that the discussion of this paper will throw more light on these points, which are generally better understood by European than by American engineers, and full and free criticism is therefore invited.

SUMMARY: This paper is an historical review of the United Fruit Company's Radio Towers, and describes the origin of a 1906 design for 200 ft. (61 m.) towers; of 50 ft. (15 m.) extension added in 1910; of the hurricane wrecks of 1914, 1915, and 1916, of data deduced from them, and of the reconstruction of the towers. It discusses the effect of a new policy that "radio communication should be made so reliable that, when all other means fail, the public can confidently look to the radio stations to maintain service, which can be done if only the towers remain standing."

The preliminary designs for three-legged towers made in 1915 are described, and their economy and advantages as compared with four-legged towers, the saving due to adjustable rod bracing and the low cost of shop work effected by making all leg sections equal short chords of a circular arc, are pointed out.

The design of the Tegucigalpa (Honduras) towers, with antenna bridges and the latest revised, "Standard Plan of 400 ft (122 m.) self-supporting towers, with 150 ft. (46 m.) antenna bridges," are shown with the specified loads, unit stresses, and notes on the most favorable direction of wire pull in relation to cross-section of tower. The design of sections and details are briefly referred to, with the limiting heights of towers to which similar sections are applicable.

Two pairs of 100 ft. (30.5 m.) broadcast station towers of the Radio Corporation of America are illustrated, and the methods adopted to distribute the loads to the building frames are described.

The economic comparison of semi-guyed towers and guyed steel masts with self-supporting towers is discussed, and the preliminary design for a tower 1,000 ft. (305 m.) high is shown.

In conclusion, the theorem is proposed that the truthful application of the laws of statics and economics will yield aesthetic results in tower design, and vice versa, and the bearing of this theorem on the appearance of the foregoing towers is given.

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DISCUSSION

Robins Fleming (by letter):* Mr. Buel has presented a valuable paper on the design of self-supporting radio towers that must withstand high wind velocities. The detail and thoroughness with which he has treated the subject make the paper one that will long be turned to for reference. Some comments may be in order, even if at times they dissent from the opinions of the author.

Since the pressure of wind varies as the square of its velocity, the pressure from a velocity of 140 miles per hour is 1.96 times that fron a velocity of 100 miles and 2.4 times that from a velocity of 90 miles. From the well-established formula $P = 0.004\ V^2$ the respective pressures on vertical surfaces from horizontal winds of 140, 100, and 90 miles per hour are 78.4, 40, and 32.4 lbs. per sq. ft. But the actual velocity is less than the indicated velocity with a corresponding reduction in pressure. (The velocities recorded and published by the United States Weather Bureau are those indicated by a Robinson anemometer, not the actual). Again, average pressure on large areas are less than those on small areas. On the other hand, the published maximum velocities are those maintained for periods of 5 minutes duration, but for shorter periods, say for a minute, these velocities may be exceeded. The writer considers the 30 lbs. per sq. ft. of exposed surface which he has recommended for towers not over 300 ft. high to be sufficient when the indicated wind velocity does not exceed 100 miles per hour. For triangular towers the area of exposed surface is assumed to be 1½ times and for square or rectangular towers two times the vertical projection of one face. For the upper parts of towers more than 300 ft. high, 2½ lbs. is added for every 100 ft. additional height. For towers within the hurricane belt these pressures would be modified, probably near to those used by Mr. Buel.

A working stress of 24,000 lbs. per sq. in. net tension is used by Mr. Buel against the 20,000 lbs. of the writer. This is not excessive for high wind velocities tho there is little left for emergencies. A decided objection, however, is raised against the use of the compression formula: $24{,}000 - 50\ l/r$. On a basis of 24,000 lbs. for tension the widely-used formula $16{,}000 - 70\ l/r$ becomes $24{,}000 - 105\ l/r$. Mr. Buel cites for justification of his use of 24,000 lbs. for tension the specifications of the American Institute of Steel Construction. According to these same speci-

*Received by the Editor September 5 1923.

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fications the formula for compression due to wind pressure would be $\frac{24{,}000}{1+l^2/18{,}000\,r^2}$. The respective values obtained by the three formulas for $l/r = 1{,}000$ are 19,000, 13,500, and 15,400; for $l/r = 140$ are 17,000, 9,300, and 11,500; for $l/r = 160$ are 16,000, 7,200, and 9,900. It is believed that the values obtained by the Buel formula are excessive and should not be used.

The discussion of square versus triangular towers will be left to others. The triangular towers weigh less, but most structural firms will quote a price for fabrication and erection that will more than offset the saving in weight, especially if the towers are of the outlines shown in Figures 23 and 27. Just here a tribute is paid to Mr. Buel for the graceful outlines he secures. Will the additional expense warrant him in so doing? His 350- and 400-ft. towers are not in New York or Boston, but at Almirante and Tegucigalpa.

Mr. Buel advocates the use of rods instead of angles for diagonals in the faces of towers. It is true he thus reduces the area exposed to wind, but he encounters serious difficulties in erection. The rods must be adjusted and he writes—"it is a very difficult job, requiring an expert, to line up a crooked tower under wind—and the wind blows a stiff breeze most of the year in the trade wind belt." From the hazardous nature of the work the only "experts" available are the workmen engaged in the erection of the towers. However skilled some of these men may become, it may be questioned whether they are ever able at a height of two or three hundred feet to tighten some rods and leave others slack so that when the wind reverses, the tower will not be thrown out of line. With angle diagonals no adjustment need be made and a tower can be kept plumb as it is erected. Angles are not as liable as rods to be bent in transportation and handling and have greater rigidity when in place.

It should also be noted that the stresses in rods are indeterminate, owing to the initial tension brought upon them in their adjustment. An initial tension of 3,000 lbs. on a rod 1 in. diameter under the thread adds a stress of 3,800 lbs. per sq. in. of sectional area. The total stress is thus brought to 27,800 lbs. per sq. in. if the rod is designed for 24,000 lbs. per sq. in. The initial tension also causes additional stresses in columns and struts.

Mr. Buel favors painted instead of galvanized towers. He states. "The cost of maintenance covering a period of fifteen years shows that frequent painting is cheaper than galvanizing."

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A difference of opinion exists on this subject. A letter, before the writer, from another designer of radio towers reads, "The cost of repainting these radio towers in the field from time to time is such that the galvanized are much preferable and more economical." As the cost of painting high towers is quite formidable there will probably be a tendency to neglect it, thus furthering corrosion.

The use of Irving Subway grating for walkway floors and galvanized pipe for compression members in suspended cantilevers is novel and unique.

The writer thoroughly agrees with the author in what he says regarding the superiority of self-supporting towers over semi-guyed towers or steel masts.

The thanks of structural engineers are due the author for including in his paper brief descriptions of the broadcast stations on Aeolian Hall, New York City and on a reinforced concrete building, Washington, D. C.

In conclusion, the behavior of the United Fruit Company's towers during the next ten years will be watched with interest.

R. D. Coombs, M.A.S.C.E. (by letter):* Mr. Buel's paper is a valuable contribution to the somewhat meagre literature on a most interesting field of engineering, as well as an expression of practices in a new part of that field—the radio tower. The general field embraces transmission line towers and radio towers, the difference between these classes being that the usual transmission tower is not a dead end structure while the radio tower is even normally a dead end for part of its load.

In this general field the designer must depart from the shelter of the usual working stress or factor of safety and work thru a more logical and professional analysis of the actually expected loads and the practicable unit stresses. He must also very clearly set forth just what load conditions are economically justified and must realize that the usual handbook unit stesses are not for him. The designer of towers may not take refuge behind an assortment of hidden factors of safety or relief on the imaginary factor of safety of four of the usual building codes.

In most steel construction the wind load is not given much active consideration. The assumed live and dead loads may be so great that the structure can carry a very considerable wind load, provided the other loads are not all present. Or with small dead and live loads the structure might simply be unable to carry

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the usual wind loads and would probably never receive them.

In towers, however, the wind load is of the most inportance both in its direct effect and as greatly increasing the pull of conductors, antennas, and so on.

Further, since the designer knows that some time the tower will receive a wind load of considerable amount, the question is as to the amount to be provided for in this event. Is it the greatest recorded pressure in that locality, or more to provide for contingencies, or less to allow for probability of occurrence some five, ten, or more years hence?

The author's examples are believed to be unique as scattered installations deliberately calculated to withstand "hurricane" pressures. The apparent procedure of determining the greatest wind reasonable for the installation in question and then providing a known, or closely known, factor of safety in the unit stress alone is to be commended.

The author's examples of failures are amongst the few service failures of larger towers of which much is known. It is unfortunate that except in tower tests the actual loads are never known because there is never any accurate measure of the wind load that was imposed on the structure.

In the writer's opinion the formula used by Mr. Buel for the ultimate stress, $44{,}000 - \frac{l}{r}$, is about as accurate for towers as indicated by the data available. The writer would, however, question Mr. Buel's bolt values, which seem to be relatively low.

It would appear that painting every two years does not indicate a very dry climate, and the writer is not prepared to admit that 1/16 inch (0.16 cm.) of steel is as good protection as a film of zinc—provided the film is continuous.

The author is requested to enlarge on the question of lock washers versus finger-loose nuts in view of the difficulty of access to such high structures.

The writer is aware of the opinion and practice of the average field crew regarding washers and that most of the transmission towers have none, but is not sure that lock washers may not be a good investment.

The use of rod bracing for such structures appeals to the writer as logical, but involves careful erection since it is very easy to build a "wind" into a tall tower. The writer's company has designed and built both rod-and angle-braced towers of medium height, about 150 feet (45.7 m.), and have believed that the rod-braced were not commercially economical for such heights. Rods

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commend themselves from a wind pressure, and maintenance painting standpoint altho thought to be noisy and difficult of erection.

Referring to the Swan Island towers, were they not relatively weak in the upper half? The parabolic outline, also used by the writer, will produce a graceful appearance and a very gradual decrease of main leg stress, so that the upper half may readily be kept below the stress of the more usually critical sections. In this connection it may be noted that the technical press recently illustrated, but without descriptive detail, a tall tower on the Rhine that indicates a reversed parabola, convex outward, and a far from graceful outline.

George A. McKay (by letter):* 1. The article prepared by Mr. Buel is a most valuable contribution to an important branch of structural engineering. As in all special branches, progress is marked by steady growth based on past experience, and Mr. Buel, in putting into print the record of his achievements, has enabled the engineering profession in general to benefit thereby.

2. The Bureau of Yards and Docks of the Navy Department, which has the responsibility for the design and construction of all radio towers for the Navy Department, started in 1912 with towers of square cross section, installing one 600-foot (183 m.) and two 450-foot (137 m.) towers at Radio, Virginia. These three towers were developed to withstand a wind pressure of 30 lbs. per square foot (147 kg. per sq. m.) over one and one-half times the projected area of one side in accordance with the usual viaduct practice, and in addition to the wind load, a pull of 10,000 lbs. (4,600 kg.) from the antennas. In May, 1913, however, when the 600-foot (183 m.) radio towers for Darien, Canal Zone, were being designed, it was decided to adopt the triangular cross section type of tower, principally on account of the economy thus effected. It will be noted from the above that the adoption of the triangular cross section by the Bureau antedated its adoption by the companies representated by Mr. Buel.

3. A number of triangular towers, ranging from 150 to 820 feet (45.7 to 250 m.) in height, have been erected in the past 11 years by this Bureau in latitudes varying from south of the equator to Alaska, and these towers have been subjected to hurricanes and ice loads. The Bureau has yet to record a single failure.

4. An interesting comparison exists in the fact that the

*Received by the Editor, September 12, 1923.

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Bureau's standard design for 200-foot (61 m.) towers weighs approximately 13 tons (11,800 kg.) on the basis of a triangular cross section, whereas the 200-foot (61 m.) square towers which collapsed, as mentioned in Mr. Buel's paper, weighed 22½ tons (20,400 kg.), or approximately one and seven-tenths times as much as the Bureau type.

5. An examination of Mr. Buel's original design, indicated on Figure 1, before the towers were strengthened, indicates that the rivets in the splices for the main column sections at Panel *P* were considerably weaker than the columns themselves, considered either in tension or compression. The method of connection of the diagonal rods to the struts is also subject to criticism. In addition to this, the use of ordinary bolts, with possible line bearing instead of bearing over the semi-periphery, would also militate strongly against the towers withstanding abnormal loadings.

6. The Bureau adopted in May, 1914, in connection with the radio towers for New Orleans, rod bracing instead of stiff bracing, in order to reduce the superficial wind area, and has retained this detail for all towers constructed since that date. It is to be noted from examination of the Bureau's drawings that the details of the connections of the diagonals to the main columns are such as to practically preclude torsional or bending stresses in riveted connections.

7. It is sincerely to be hoped that the publication of the article by Mr. Buel, with comment by various engineers thruout the country, will give designing engineers, generally, the necessary data to develop satisfactory designs for future towers which the progress of radio communication will demand.

Schuyler B. Knox, Member American Society Civil Engineers (by letter):* The fixed, triangular type of radio tower construction has advantages over a greater-sided system in its better resistance to deformation thruout its height and the saving in weight of steel by the elimination among other things of at least a column leg and the horizontal sway-bracing at panel points. Also one should not overlook the less number of masonry pedestals required for column leg footings.

In the case of high structures designed to carry wind loads of great intensity even in our temperate zone, the designer is to be commended for his employment of tension rods on which the exposed wind surface is one-half, or less than one-half, that

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of angle diagonals of equivalent cross-sectional area. The materials used in the "bridge" construction at top of the tower also present a minimum surface exposed to wind action.

By reducing exposed surfaces in the very high radio towers in localities subject to violent hurricanes, Mr. Buel has solved the problem economically, for in this way he has lessened the amount of steel required to resist the intensity of wind stresses in legs and web members without changing the strength of the structure or causing additional cost in the aggregate over square and other types. The higher unit cost of fabricating due to adjustable, upset rods, including pins, clevises and "safety" set-screws is generally offset by the smaller total tonnage in the triangular type structure.

In the fabrication of the 350-ft. (107-m.) triangular towers for the United Fruit Company's Almirante Station, Panama, for which the Fort Pitt Bridge Works (with which the undersigned is connected) had contract in 1920, no difficulty was experienced in the shops, due to co-operation between the engineers and the fabricator in the matter of details and determining methods for handling. Both towers were assembled at the shop, lined up with transit, adjusted, and match-marked while so assembled. In this position the adjustable rod diagonals were fixed in length by keying the clevises with "safety" set screws so that no variation would be possible after the work was knocked down, thus insuring greater accuracy in plumbing during erection. The members were then separated and packed for shipment.

Quoting from Mr. Buel's report of one of those conferences: "Since tensile and compressive stresses in legs are nearly equal, and as bolts at splices must take all tension and as end bearing is unreliable owing to flexure of tower and angles and might set up very objectionable eccentric or bending stresses in the leg sections, it was agreed to put 100 percent field bolts in all splices and leave joints ¼ inch (0.64 cm.) open—same as established practice with hip joint in bridge chords."

While the fabrication costs per unit weight of steel may be higher than for other forms of towers, the saving in the number of masonry footings, the less number of, and lighter weight of members, not to mention the lightness and appropriateness of the "bridge" at the top of the towers, tend towards making the total cost of installing this tower, from a purchaser's point of view, as low as, if not lower, than for many other forms. Mr. Buel's design is scientifically developed.

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Reginald A. Fessenden (by letter): Mr. Buel's paper is very interesting and valuable, and puts radio tower construction on a sound footing.

The three-legged tower with pin connection is undoubtedly the right type. Anything which concentrates the material into few and large pieces minimizes the effect of corrosion and saves a great deal more steel in the end, and adds greatly to reliability, which is, of course, the main thing. And with the wind whipping the towers about the way it does, there is great danger, in a four-legged tower, of getting very concentrated stresses at certain points, which will not occur with a three-legged pin-connected tower.

About corrosion, two things occur to me. First, what would be the comparative costs of one of the new non-corrosive chrome steels compared with standard material? Possibly much too high; yet a comparison would be interesting.

Second, what would be the effect of always keeping the tower hydrogen polarized to ground? Possibly nothing; and yet I have found that corrosion increases very much faster when oxygen polarization is superimposed upon alternating current flow, and of course one gets this alternating current flow from the radio sending. So there might be some good effect with hydrogen polarization.

The curved legs are a fine idea, and it is well worth knowing that it can be carried out in practice with so little expense.

Can Mr. Buel tell us if there would not be some advantage in using isolated guy wires between the towers themselves, so all would act as a unit. I note he refers to the loss from eddy currents in the towers themselves and in long guys, but this can be got over very easily by slipping very small coils of a few turns of soft iron or nickel steel wire over the guys or tower members. This has the same effect as putting in insulators, so far as eddy currents are concerned, and without any of the expense or bother. See United States patent 706,746, "Wave Chute," August 12, 1902.

The fact that the foundation can settle with the three-legged type without hurting anything is very important, and they are also better for twisting stresses. In this connection, any concrete foundation pillars should have the reinforcement arranged for twisting stresses, as these may be very bad in earthquakes. Another good point which strikes me is the adjustable tension rods.

About the wind stresses, has any one tried making these

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measurements on different shaped elements while mounted on an aeroplane? It would look as if this would be a very simple way, and the models or full-sized cross section members could be made of wood. On the Brant Rock tower I used a fine wire of gold, with one-half of it mounted in a thin metal tube and the other in the wind, and a tap from the center, and found how much current must be passed thru the outside half to make it the same resistance as the shielded half. This gave me the instantaneous wind velocity, as it responded in about one-tenth of a second. If made recording, this might be useful for getting maximum gust velocities.

I certainly like the appearance of the towers. They look like something. It will not be long before we have one million volt transmission lines, and Mr. Buel has shown us that it is possible to get something which will not disfigure the landscape.

Jacob Feld, 1077 East Twelfth Street, Brooklyn, New York (by letter):* The author is to be congratulated for his very clear and instructive presentation of a comparatively recent type of structure which the engineer is called upon to design. Much more can be learned from a description of a failure with a careful analysis of the causes than from the usual account of a successful (or rather, untested) structure. From this point of view, this paper is especially valuable.

The predominating characteristic of self-contained radio towers are their flexibility. As Mr. Buel has pointed out, the deformation of the tower relieves the tension in the antenna by decreasing the chord length of the antenna wires under wind-load. It is noteworthy that there is only an increase of 17 per cent over the dead load tension, when the antenna is exposed to maximum wind. This requires considerable deflection of the towers, introducing a rather complicated stress distribution in the legs of the tower. The relieving effect of the towers permit considerable sag of the antenna wires (in the direction of the resultant of wind and dead load). The changed geometrical shape of the antenna has some effect upon the electrical characteristics of the system. The writer hopes that the discussion will bring out an explanation of the magnitude of this effect, and what limits are permissible. It seems to the writer that too much deflection causes trouble electrically as well as structurally.

The description of the mast shown in Figure 5 is especially instructive and emphasizes some of the usual rules for economical

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design of structures. The mast was strengthened by the addition of two guys placed at 30° to the vertical. It would have been more economical to have increased this angle, requiring, of course, a longer guy and more room, but decreasing the effect of the guy upon the tower and increasing the beneficial effect of the guy in taking wind loads. The tower failed because of the absence of windward supports after the antenna broke from the mast. This points out an important feature—the necessity of symmetry in supporting a tower subjected to a wind force from any direction and the necessity of observing the fundamental principle of making each unit of a structure self-supporting.

In the case under consideration, Mr. Buel states that no guys were permitted inside the quadrant covered by the antenna. That did not prevent the placing of a system of four guys at 90° to each other, or three guys at 120° to each other. In either way, wind from any direction can be taken care of. A careful analysis, too lengthy for this discussion, will show that the three-guy system is more economical, as far as weight is concerned, but not so rigid as the four-guy system. A tower or mast with three main legs, or triangular in section, naturally lends itself to simple connection details for the three guys; similarly for the four guys with a square tower. The writer was very much astonished to see the description of the tower which failed after the antenna gave way. It is the basic idea of design to make, in as far as possible, each section of a structure self-dependent for its stability. It seems that a radio system should consist of a stable tower or mast, which in addition to being capable of holding up itself and any loads which may act upon it, can also resist certain external forces caused by the antenna.

An almost analogous case is the usual bridge design: a substructure, self-supporting, with additional stability to hold up the superstructure. Certainly, no one would think of designing a bridge so that, should the superstructure fail, the substructure was no longer stable.

Mr. Buel describes how a tower was increased in height and exposed to a greater antenna pull without strengthening and without informing the designer. Unfortunately, this state of affairs is very common. The radio engineer must remember that a design according to specifications, in order that it be economical, is made to withstand safely the loads specified, and no more. If the radio engineer continues to increase the loads upon the towers, failures are certain to occur. Perhaps some

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simple method of rating the towers, according to their excess stability, will help prevent the overloading of towers.

It is interesting to note that the towers failed in the panel just below those which had been reinforced by the substitution of "K" bracing for the flexible tie-rod bracing. This is an example of the very frequent error made in strengthening a structure by increasing the stiffness of assumed or tested weak points, thereby making certain isolated portions more rigid than the rest. An accurate analysis is tedious, but it may be stated without computation that the transmission of stresses across the section between two portions of unequal rigidity causes some extremely localized summations of stress. This has been very clearly pointed out in the recent researches on the transmission of stresses along cracked and punctured members. Sudden changes in section of the tower as a unit, as well as of the individual legs and secondary members must be avoided. This is especially true for high members affected by relatively high bending and twisting moments.

The author is to be congratulated for his clear description of the advantages of the three- over the four-leg towers. The writer questions the advisibility of using bent plates and curved legs with such high unit stresses, 24,000 pounds for tension and $24{,}000 - 50\frac{l}{r}$ for compression. Even tho the torsion is taken into account, some allowance must be made for the secondary stress and the reversal of stress. The writer is fully aware of the high unit stresses under which transmission towers have failed when tested to destruction. The only justification for such high unit stresses is the rather high unit pressures assumed for the specified wind velocities. Mr. Buel uses a conversion formula, the relation between wind pressures and velocity, which is probably at least 20 percent excessive.

The use of rods for wind bracing in place of stiffer sections is certainly desirable, because it eliminates so much wind load. It must be remembered, however, that one stiff diagonal will replace two rods in each panel, so that the saving is not so large as it seems at first glance. The main advantage is in the smaller exposed wind area, especially since the effect on a cylindrical rod is less than on a flat surface. The writer believes that we can go one step further and, where possible, use steel wire rope for the tension members. It would hardly be possible to use wire strands for the bracing in the towers, but it certainly can be used for the ties connecting the ends of the antenna bridge

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with the mast. Initial tension can easily be provided for stiffness, and, in place of the upset rods which are liable to bend and twist under high wind, a lighter and stronger steel strand can be used. The saving may be small if figured by weight, but the leverage is large and every square inch of exposed area at that height counts. The chief advantages of the wire rope are the high allowable unit stress and the safety from bends.

With such small ratios of base width to height, it is especially important that there be no unequal settlement of the footings. The writer would like to know whether Mr. Buel has noticed any trouble from this source and how he prevents it.

The author's statement that the maximum stress in the antenna can be controlled by attaching the antenna to a counterweight is noteworthy. Such devices have been used in several tower designs, especially those of foreign types. This method relieves some of the antenna pull upon the tower and decreases the deflection under wind load, but it increases the vertical load upon the tower.

It seems to the writer that Mr. Buel has not fully appreciated the value of guys. With the high unit stresses allowed, they form most efficient tension members, both from the point of view of economy in weight and because they are less affected by wind. Not only are they thinner than the stiff members, but the unit pressure of wind upon a circular rope or rod is also less than the unit pressure upon a flat surface. Mr. Buel's objection to guys on the ground that they do not last is not substantiated by experience with wire cables and strands in suspension bridges. The suspenders of the Brooklyn Bridge are in very good condition. The wire ropes removed from the old Niagara suspension, after 50 years of service, were remarkably unchanged. As far as maintenance is concerned, Mr. Buel points out that the steel of the towers must be painted on the average of every two years. Of course the use of guys necessitates land areas, but radio stations of great heights are seldom located in districts where land is expensive, and it will be found that the percentage of increase in the land required is small, even tho the actual area may seem large.

Albert W. Buel (by letter):* Mr. Coombs questions the bolt values used for the designs shown, which, he says, "seem relatively low." These were adopted to permit the use of rough bolts, not turned, and also to provide for the reversals of stress—the ten-

*Consulting Engineer, 29 Broadway, New York. Received December 15, 1923.

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sion running as high as 90 percent of the compression in some leg sections.

Spring lock washers have given some trouble by breaking and dropping out. Moreover, it is thought that the plates will be gripped more tightly without them and, after all, the frictional grip should not be neglected. The nuts are gone over, tightened, and checked with a chisel after erection and periodically thereafter. Very little trouble has been reported.

Rod bracing will not show much economy in towers much under 200 feet (61 m.) in height, but for the higher towers the saving is as much as 20 percent in weight. By adjusting the rods at the shops the towers can readily be erected out of wind.

The Swan Island towers were weak in the upper half, particularly between O and R, about 100 feet (31 m.) below the top, due to the insertion of the 50 ft. (15 m.) extension, T-10 to T in Figure 16, without reinforcing the members below. There was originally some excess section in the legs at some stories, but this is often unavoidable in light compression members where practical considerations are given due weight.

Mr. Fleming says the tensile unit stress used for the later designs (24,000 lbs. per square inch, against his 20,000 lbs.) "is not excessive for high wind velocities." It is not clear to the author what wind velocities have to do with unit stresses, nor why 24,000 lbs. per square inch is not just as good where the loads and stresses are computed for 90 or 100 miles per hour wind velocity (if such is the maximum anticipated for the locality), as where hurricane velocities have to be provided for. By using the maximum probable loads we make reasonable provision for "emergencies," as far as is economically justified, and avoid the confusion of loads with unit stresses. Then and only then can unit stresses be considered on their own merit.

The compressive unit stress used ($24{,}000 - 50\ l/r$) is two-thirds the ultimate strength of long columns as shown by the studies that have been made of the most recent tests. ("Transactions American Civil Engineers, volume LXXXIII, pages 1583 to 1688 and 1960.) While the results of these tests showed variations of 15 to 20 percent above and below the mean values, it is not thought that such extreme values are truly representative of these tower legs, and that the Tetmajer formula ($44{,}000 - 162\ l/r$), which gives 27,800 pounds per square inch for $l/r - 100$, is nearer the correct value for these sections.

The higher values of l/r mentioned by Mr. Fleming occur only in struts where the sections are not determined by stress,

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but by limiting values of l/r, and the column formula does not apply.

Mr. Fleming says of three-legged towers: "most structural firms will quote a price for fabrication and erection that will more than offset the saving in weight, especially if the towers are of the outlines shown in Figures 23 and 27." The only additional item of cost of the three-legged feature is the 15 degrees bend in the gusset plates, of which each tower of Figure 27 has 144. About 200 such plates can be bent cold on a bending machine in eight hours, at an outside cost of $50.00, or $36.00 per tower. Each tower weighs 246,000 pounds, while a four-legged tower on the same specifications would weigh about 25,000 pounds more, at an additional cost for "freight or boat" shops of about $1,750.00, showing a net saving of $1,714.00 for the three-legged feature, without considering freight and erection. Freight charges would be about 10 percent more for the four-legged tower, and, in erection, there would be about one-third more members to handle and set and connections to make than there are in the three-legged tower. Erection equipment has been worked out avoiding interference with sub-struts, and the like.

If by "outlines" Mr. Fleming refers to the curve, of which the leg members are short chords, the only cost of this feature is that the gusset plates require a special lay-out and template at each story. There are six duplicate plates of each pattern in a tower and the last order was for eight towers, giving 48 duplicate plates from each template. Surely, with such duplication, the extra cost of layouts and templates could not much exceed $50.00 per tower.

Mr. Fleming intimates that appearance is wasted in "Almirante and Tegucigalpa," but might be warranted in "New York or Boston." On the contrary, the author is inclined to reverse that idea where the countries are of Latin extraction and civilization. But it should be noted that the curved outline has real and practical advantages on the side of economy which determined its adoption, and that its appearance was only a by-product.

The difficulty of adjustment of rods in the field is overcome by setting them to calculated lengths and locking them with safety set-screws at the shops. Very little adjustment has to be done after they are erected in place. The initial tension in rods does not act to increase the total maximum stress unless it is greater than the stress from loads.

Mr. McKay says that "the Bureau (of Yards and Docks) has

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yet to record a single failure," and his next paragraph sets up a comparison which is misleading and not based on the facts stated in the paper. No failure has yet occurred in any tower designed by the author except where the 50 foot extension was added by others and the wire pull increased from 2,000 to 7,000 pounds, without increasing or reinforcing the sections below, and without consulting the designer. Excluding the shoes and anchors, the original towers weighed only 20 tons. Mr. McKay gives the estimated weight of the Bureau's 184-foot design for comparison with the original 200-ft. design described in the paper. Moreover, while the Bureau's design specifies "minimum thickness of material 5/16 inch. except as shown," the webs of the channels composing the legs for the lower 178 feet of the 300-foot design are under ¼ inch thick—under 7/32 inch for two stories—and the same is true for the main struts up to 116 feet above the base, and the main connections are on these thin webs. Such sections are theoretically economical and permit much longer unsupported lengths with fewer struts and diagonals, showing a very large saving in weight, but leave small margin for deterioration by corrosion. Such thin webs are weak links which, in the author's opinion, should be used only for temporary work or where conditions are most favorable for protection. Laced channel columns, with flanges turned in, are much more expensive to fabricate than the sections shown in the paper, and, as has been noted, final costs and nor weights alone should be compared to determine relative economy. The original 200-ft. design had single angle legs, so much cheaper to fabricate compared with the sections used by the Bureau, that weight alone loses significance. The design was modified in some details to suit a certain shop with limited facilities, to which the order was allotted, which explains the economical diagonal hitch used. It has served its purpose as most of the original towers are still in service, but in all later designs an improved connection has been used, with small eccentricity and no tension on bolts or rivet heads.

Figure 1 shows the top of tower four feet square, providing for considerable resistance to torsion, but adding considerably to weight and cost as compared with the Bureau's design with its top only about seven inches on centers of gravity of the angles and negligible resistance to torsion. Where loads, specifications and details are so different, comparisons of weights are likely to be misleading.

The author has made designs for towers 600 ft. and 1,000 ft.

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high, for loadings similar to those used by the Bureau, from which the interpolated weight of an 820-ft. tower is only 350 tons as against 560 tons, the estimated weight of the Lafayette Station towers on the Bureau's design, as given by Mr. D. Graham Copeland. This shows a saving in weight of 37.5 percent with the Bureau's design.

Estimated weights from the design in Figure 19 are, for a 240-ft. tower, 26 tons, and for a 300-ft. tower, 37.5 tons, which give when reduced to the Bureau's loadings, 17.25 tons and 25 tons, respectively, as against 19 tons and 30 tons from the Bureau's design. This indicates a saving of 10 to 16 percent, notwithstanding Figure 19 shows a top width of seven feet compared with the seven inches of the Bureau's design.

The photographs of the wrecks show that all splices and connections held, confirming out judgment that the bolted connections were properly proportioned. Practically the bolts by friction, and theoretical bearing and shear do not act until slipping takes place. When and if bearing comes into play, the theoretical line contact is enlarged, elastically, as has been conclusively shown in the case of locomotive drivers on rails.

Commenting on the discussion by Mr. Knox, credit for the method of locking the adjustable rods with safety set screws and for a number of tower details is due to the Fort Pitt Bridge Works.

All details and connections of the walkway grating on the bridge were developed by the Irving Iron Works Company.

Professor Fessenden asks for comparative costs of non-corrosive alloy steels. About two years ago the author made some inquiries along this line with discouraging results. The tonnage of single orders has not yet been sufficient to warrant the use of alloy steels with the delay in deliveries which would probably ensue. These preliminary inquiries indicated that the cost would be excessive.

The nearest approach to a system of towers guyed to each other, "so that all would act as a unit," are the four Swan Island towers connected by two messenger cables and the antennas—see Figure 5. Unless outside guys are added, as at Swan Island, the connection between the towers is of small advantage, as the windward tower receives no support. The paragraphs on semi-guyed towers seems to apply to this case.

Professor Fessenden's suggestion that concrete pedestals should be reinforced for the twisting stresses due to earthquakes, deserves careful consideration. The pedestals of Las Vacas

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Viaduct went thru the four severe shocks and many minor shocks of December, 1917, and January, 1918, which were of several different types, without any reported injury.

The Bureau of Standards has made many tests or experiments on wind pressure in wind tunnels, which, altho not directly applicable in cases of tower design, are instructive and helpful.

Referring to Mr. Feld's discussion, it is important to note that only with messenger cable or triatic systems is the wind increment in wire pull as low as 17 percent of the dead load pull, and tower deflections are not unprecedently large. If "rather complicated stress distribution in the legs" refers to secondary stresses due to deflection, the author considers that they are considerably less than in cantilever bridges, and a remark made in a similar case by that eminent bridge engineer, the late Paul L. Wolfel, "it is an elastic structure and will take care of itself," seems applicable. Under wind loads the antennas blow out of line, which may be objectionable, but the maximum sag is fixed at the minimum at which the strands will carry the loads. A "T" system, with anchored downleads, holds them nearer in line. The economical angle of guys—45 degrees—compared with that of Figure 5, would not have saved over $110.00 for the four towers, and other considerations favored an angle of about 30 degrees. At that time guys were not considered permissible except in the outside quadrant, and not near the mat rectangle. They were used only as an expedient to salvage the undamaged parts of the towers, and six years of service, during which they resisted a severe hurricane, has justified their use. Considering the history of Swan Island Station, it is not quite clear to the author at just what Mr. Feld was "astonished."

Mr. Feld is right as to the effect of the "K" bracing in story P-Q in principal, but they were additional reinforcement—not a "substitution," and it is important to note that they did not tie into the upper intersection at "Q," but left an unsupported length of 3′ 6″ for the 4″ leg angle. Also that stories M-N-O-P and Q-R-S-T were nearly as weak and, after P-Q was reinforced, they were the critical ones. As one of the towers failed in story Q-R, above the "K" bracing, the influence of the latter could not have been very great except to prevent failure from occurring in story P-Q. Quite likely this was sufficient to cause two towers to fail in O-P instead of in M-N, as might have been expected. (Since they were generally reinforced from G to T, they have behaved very satisfactorily in a severe hurricane.) The only importance this now has is the effect on the author's computa-

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tions of probable average wind pressure at the instant of failure due to neglect of secondary stresses. but this error is on the side of safety and less than the limit of accuracy assumed for the method. The wind pressure adopted from these studies is considered safe, but not excessive for that locality.

Mr. Feld says "one stiff diagonal will replace two rods," but the slenderness ratio would then require sections for diagonals heavier than the horizontal struts, excluding a single system of bracing with reverse stresses from serious consideration. The substitution of wire rope for tension members, which he suggests, would require the fabrication to be split up between structural shops and wire mills. Even with pin-connected, adjustable rods and pipe frames for antenna bridges, several large fabricators have not considered the work very desirable, and it is considered preferable to simplify and standardize rather than to introduce new complications which might prove very costly, altho invitations have not yet failed to get several satisfactory proposals. Inquiries for steel guy lines, protected and wrapped like the cables of large suspension bridges, have not found any manufacturer equipped to supply them. Increases in vertical loads within reasonable limits would not cause proportional increases in total stresses, as the wind stresses are the larger component, and vertical loads increase resistance to overturning.

Mr. Feld says "it is especially important that there be no unequal settlement of the footings." With four-legged towers such settlements would be serious, but with three-legged towers the only effect would be to throw the tower out of plumb, and very considerable inequalities in settlement of the three pedestals would have to occur before the stability would be in danger or the stresses dangerously increased. On the other hand, a very small inequality in settlement of a pedestal of a four-legged tower might increase stresses to the point of failure, especially in the diagonals.

The author thanks all contributors to the discussion, especially for analytical and critical comments, which have helped to clear up some points and to direct attention to others that might have remained obscure.

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Further Discussion on "An Improved Systfm of Modulation in Radio Telephony*

By Charles A. Culver

R. A. Heising (by letter):† In looking over Mr. Culver's paper entitled "An Improved System of Modulation In Radio Telephony," I find a few points on which I differ from him. The system which he describes is not new to us because of its having been previously disclosed by Lockwood and Beauvais, as he mentioned, and also because it was independently invented by Mr. L. J. Sivian of this Company in 1917 or 1918. At that time, Mr. Sivian made a study of the system and his results did not show the system to be quite such a favorable one as Mr. Culver claims. Mr. Sivian tried out, not only the specific form which Culver showed, but numerous modifications of it. The best one of these modifications compared favorably with the constant current system as regards efficiency and completeness of modulation, but in some other respects it was inferior. Its greatest defect was that it had a curved control curve which would indicate that considerable distortion would occur in process of modulation. This curvature was such as to give distortion considerably greater than that in the constant current system. The system also would work with only certain tubes. One requirement was a high amplification constant. It also appeared to work better with tungsten filament tubes, that is, with tubes which did not operate with a wide margin of temperature saturation. In many of the circuits which he tried, difficulty in adjustment was encountered. In order to make the control curve as straight as possible, specific adjustments had to be made which made it more difficult to operate or manipulate than some other circuits. He felt, at first, that he had found a system which was superior to any others, as has Mr. Culver, but after an exhaustive

* See Proceedings of The Institute of Radio Engineers, volume 11, number 5, October, 1923.
† Received by the Editor, September 12, 1923.

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study of the subject, he came to the conclusion that, in almost every respect, it was considerably inferior to what we were using.

As regards modulation in general, there are a few points which I think would be worth while to emphasize. One of these is that whatever the modulation system, the over-all efficiency depends largely upon the efficiency of the generator of radio frequency power. A study of this efficiency over the power range that the generator operates during the modulation cycle will usually show where any shortcomings lie. If we do this for a system in which the antenna resistance is varied to do the modulating, we find that the efficiency drops to such low values and for such periods as to render such a modulation system no more efficient, and usually less so, than our present types. This is particularly true in vacuum tube oscillator or amplifier generators. The efficiency of any vacuum tube generator is high only when the impedance attached to the plate circuit is greater than the output impedance of the plate circuit. Varying the antenna circuit resistance will cause this attached impedance to vary. Since reasonable modulation cannot be secured unless the antenna resistance varies over a wide range, the attached impedance will vary over a corresponding range. It happens that reasonable modulation will not be obtained unless such a range extends on both sides of that value which corresponds to the condition for delivering maximum power and it will be found that on one side the efficiency will be so low as to result in a lowered over-all efficiency of the system.

For high efficiency in a vacuum tube generator, it is not sufficient to maintain the impedance of the attached circuit at a value equal to or greater than that of the tube, but the tube must also be operated at approximately full power. The reason for this is easily deductible from the principles pointed out by Professor J. H. Morecroft in his "American Institute of Electrical Engineers" paper of 1919. He shows that the losses within a power tube can be kept low only when the alternating potential of the plate approaches the direct potential in magnitude. Therefore when a vacuum tube is operated in any manner whatsoever so as to reduce the ratio of alternating voltage on the plate to direct voltage, the efficiency will drop. Any method of grid control or antenna resistance variation cannot but help reduce this ratio, and the efficiency of the generator will fall to such a value that the over-all efficiency of the set will, at the best, be the same as that of the constant current system, if not falling much below. This element in operation is fundamental in Mr. Culver's circuit,

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and, reasoning from that premise, I would say, without ever trying out the circuit, that its over-all efficiency could not be greater than that of existing systems.

This leaves, of course, only two ways for securing high efficiency in the generation of modulated power. One is always to operate the vacuum tube at full power, keep the impedance of the attached circuit at a large and fixed value and to do the modulating by varying the plate voltage. This keeps the efficiency of the generator up during almost the entire modulating cycle, especially during that part when very large power is being delivered. It is the method which is made use of in the constant current system. The other method is to vary, in some way or another, the coupling between the vacuum tube and the tuned circuit, keeping at all times the coupling above the amount necessary to give maximum power. That is, the coupling should be varied between the limits of the maximum power value and infinity. The variation of any constant in the generator circuit other than the coupling between these limits, will cause a reduction in efficiency during part of the cycle which will reduce the over-all efficiency to an undesired low value.

Another point on which I do not agree with Mr. Culver is the magnitude of the non-signaling current. The strength of a received signal is dependent, in the average detector, upon the magnitude of the variation of the antenna current and not upon the non-signaling current at all. It makes no difference whether the non-signaling current is A or $2A$ as Mr. Culver mentions. In the constant current system the non-signaling current is A and the variation is $2A$. His method of operation is to make the non-signaling value $2A$ and the variation $2A$. By making the non-signaling value $2A$, he does not contribute in any way to the loudness of the signal, but does waste more power in the oscillator tubes than it is necessary to do.

In any kind of a system the tubes must be built to produce an antenna current of $2A$ amplitude. If a given number of tubes are provided and all tubes are used as oscillators, we can assume that at the voltage they are operated at, say 1,500 volts, they will give an antenna current of $2A$. If this adjustment is for maximum power, than there is no possible method of control operating upon the oscillator circuit (other than plate voltage control), which will cause these tubes to give more than $2A$ current, and the modulation must be entirely downward, which is the way I understand Mr. Culver intends to operate. If, however, we should take one-half of these vacuum tubes and adjust them

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for maximum power on the same voltage, this number of tubes will give one-half the power that all would give, but half the power would be 0.7, the current that all would give. Under ideal operating conditions with one-half as oscillators in the constant current system and the other half as modulators, the non-signaling value of the current would, therefore, be 0.7 of 2*A* and the maximum current would approach twice that or 1.4 of 2*A* or 2.8*A*. This indicates that at a given voltage with a given number of vacuum tubes, the constant current system can produce in the antenna a greater variation in antenna current than where all are used as oscillators. In order to get the same power into the antenna with all tubes as oscillators, it will be necessary to increase the voltage on them by a factor between 1 and 1.4 or we would have to raise the voltage nearly to 2,100 volts, and then in order to prevent over-heating of the tubes we would have to adjust our modulating device so that the non-signaling current would not be 2.8*A*, but would be 1.4 *A*, the same as with the constant current circuit, and we find ourselves with no advantages but with a disadvantage of requiring a higher plate voltage.

In comparing his set with some other set, Mr. Culver ties his comparison to the fact that the non-signaling power to the antenna in all these cases is supposed to be 500 watts. In the sets manufactured by the Western Electric Company to give a non-signaling value of 500 watts, approximately 1,000 watts are delivered to the oscillators, 450 to the modulators, and something under 200 to all of the speech amplifiers. This, of course, includes filament currents. The oscillators deliver a steady power of 500 watts to the antenna, but under the condition of maximum modulation, the radio frequency power reaches a value close to 2,000 watts. In Mr. Culver's arrangement, the 500 watts is the maximum delivered to the antenna under any condition, so that as a matter of fact, the two sets should be compared on the basis of 2,000 to 500 rather than considering them equal power sets. Instead of showing his system to be an improvement, the total power consumption would indicate the reverse. Actually, however, the system which he describes can be adjusted so that it would not show his system up to as great a disadvantage as the preceding would lead one to believe. Under the very best conditions, he could secure equivalent power efficiency, but would have the disadvantage of having somewhat more distortion as well as requiring a higher plate voltage if he used tubes having the same constants.

Quoting distances over which a set or a system has been heard

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gives no reliable information regarding the efficiency of a modulation system. Many people, for example, in the vicinity of New York write that they listen regularly to Havana, Denver, Iowa and other distant places. However, the amount of power necessary to span such distances at broadcasting wave lengths varies thousands of times, depending on atmospheric conditions. In the case of amateur reception we have added the large variations in receiving conditions, receiving apparatus, and the personal equations of the listeners. The only satisfactory comparative test of modulation systems is successively to operate similarly power-rated sets embodying the systems, on a real or dummy antenna, measure the radio frequency power and the degree of modulation of each, in which case the results should be exactly in proportion to the product of these two factors.

This does not mean, however, that the system Mr. Culver describes is of no use. Except for the disadvantages, which in many cases are not very serious, it has much to commend it. It will give quality good enough for commercial communication. It has an "apparent" simplicity which lends itself readily to helping sales. It does not require as much amplification of the speech energy as other satisfactory systems. It is a practicable system. However, I think Mr. Culver has given it too broad a name. If he called it instead "An Improved System of Grid Circuit Modulation in Radio Telephony," there could be no question on that point.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

ISSUED OCTOBER 30, 1923–DECEMBER 18, 1923

BY

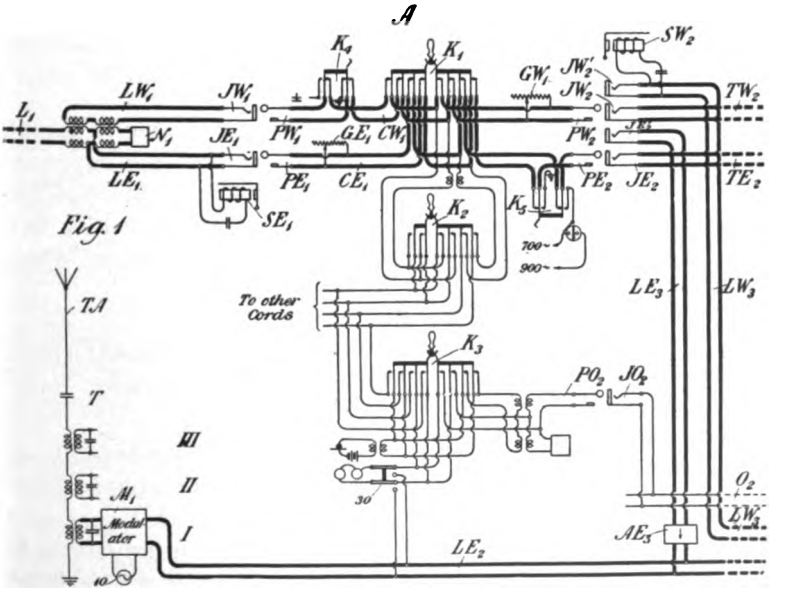
JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, DISTRICT OF COLUMBIA)

1,472,092—J. H. Round, filed July 9, 1921, issued October 30, 1923. Assigned to Radio Corporation of America.

RECEIVER FOR RADIO TELEGRAPHY, with a circuit for eliminating atmospheric disturbances, consisting of an antenna system which is slightly out of tune with the signal waves and a rectifying circuit connected with the antenna which rectifies the resultant current due to the forced and free oscillations in the antenna system. A circuit is then employed which is resonant to the beat frequency and the resulting signal is employed to actuate a responsive device.

1,472,289—R. Bown et al, filed April 26, 1921, issued October 30, 1923. Assigned to American Telephone and Telegraph Company.



NUMBER 1,472,289—Radio-Wire Connecting Circuits

*Received by the Editor, January 12, 1924.

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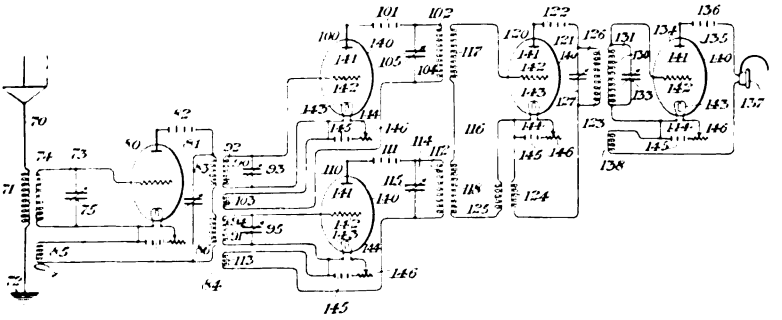
Radio-wire Connecting Circuits, in which existing line wire systems are made to function simultaneously with radio transmission and reception systems. Link circuits are provided and so organized as to interconnect wire lines with radio channels. It is necessary that the two-wire toll line be brought up to the radio station and be split into a four-wire circuit before connection with the radio transmitting and receiving channels. There are two possible places in the circuit where the necessary switching and operating arrangements may be introduced. These are in the two-wire line or in the four-wire part of the circuit.

1,472,218—J. H. Hammond, Jr., filed August 5, 1919, issued October 30, 1923.

Fig. 2



Number 1,472,218—Transmission and Receiving System

Transmission and Receiving System, in which signals are transmitted on a series of waves upon which are impressed a plurality of series of periodic modifications of different frequencies. A series of irregular modifications corresponding to a message are then impressed on said waves and modifications, which are then received with extreme selectivity.

1,472,470—R. V. L. Hartley, filed March 30, 1918, issued October 30, 1923. Assigned to Western Electric Company, Incorporated.

Method of and Means for Producing Alternating Currents by use of an oscillating electron tube having input and output circuits inductively coupled to each other with a condenser connecting the electrically remote terminals of the inductance to form with the inductances a closed oscillation circuit, which circuit determines the frequency of the oscillations produced by the

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oscillator. Another tube may be associated with the oscillator to impose a limiting effect upon the amplitude of the oscillations so that they will be the same during both positive and negative portions of the cycle.

1,472,477—R. W. King, filed August 14, 1919, issued October 30, 1923. Assigned to Western Electric Company, New York.

ELECTRON DISCHARGE DEVICE, in which a supporting standard for the electrodes is located centrally in the tube. The supporting standard is tubular in form and projects inwardly of the tube. The grid, plate, and filament are substantially mounted within the tube upon the inwardly projecting stem.

1,472,583—W. G. Cady, filed May 28, 1921, issued October 30, 1923.

METHOD OF MAINTAINING ELECTRIC CURRENTS OF CONSTANT FREQUENCY in an electron tube generator which consists in connecting a piezo-electri. resonator having two pairs of coatings with one pair connected to the output circuit and the other pair connected to the input circuit, so as to cause an alternating current to flow in the output circuit, the frequency of which is determined by the mechanical vibrations of the piezo-electric resonator. This resonator has been previously described in these columns with reference to patent 1,450,246, and it consists in general of a plate of a piezo-electric crystal with coatings on its opposite faces.

1,472,822—H. A. Affel, filed September 24, 1919, issued November 6, 1923. American Telephone and Telegraph Company of New York.

CALLING ARRANGEMENT FOR RADIO SYSTEMS, in which the usual voice currents are impressed upon a modulator at the transmitter, causing radiation of energy in accordance with speech, and wherein an alternating ringing current may also be impressed upon the modulator for radiation by the antenna to operate a ringing signal at the distant receiver.

1,472,987—P. B. Murphy, filed August 28, 1920, issued November 6, 1923. Assigned to Western Electric Company, Incorporated.

SIGNALING SYSTEM for transmitting a ringing current from a radio transmitting station to a distant radio receiver and

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causing a bell signal to be operated at the receiver. A carrier wave transmission system is provided at its transmitting station with means to cut off the local talking circuit from the outgoing carrier wave channel and to connect in its stead to the ourgoing channel a source for alternately applying two alternating currents of different frequencies in a regular sequence. These alternately applied currents are caused to modulate the outgoing carrier wave in the same manner that it is modulated by special currents when carrier wave telephony is in progress. At the receiving station, the modulated carrier wave is detected or demodulated by the same apparatus used for detection of speech modulated waves and the detected alternate frequency currents are caused to energize selective circuits to control the operation of a differentially acting relay, which in turn operates the local signal element or applies the ringing signal to the receiving circuits as the case may be. By making the receiving apparatus responsive only to the conjoint action of the predetermined frequencies applied in a predetermined manner, the possibility of false signals being received is remote.

1,473,070—S. T. Woodhull and G. T. Waller, filed September 15, 1919, issued November 6, 1923. American Radio and Research Corporation.

Gap for radio transmitters of the spark type. The patent covers construction of quenched spark gap which consists in a series of thin metal plates of uniform thickness each having a centrally disposed sparking area and an annular recess surrounding the area. Separate lengths of conducting and non-conducting material are provided which engage the front and back faces of the plates, respectively. An insulated gasket is provided between the plates forming a substantially air-tight sparking chamber.

1,473,179—R. A. Fessenden, filed November 10, 1920, issued November 6, 1923. Assigned to Submarine Signal Company, Portland, Maine.

Method for Eliminating Undesired Impulses at a receiving station, which consists in providing several channels in the antenna circuit for receiving the disturbing noises more effectively than the signals on one portion of the receiving system, recording the indications as received, and receiving said disturbing noises and desired signals in a more equally effective manner on another

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portion of said receiving system, and recording the indications so received, and detecting and eliminating the disturbing effects by comparison of said records.

1,473,220—H. F. Elliott, filed August 23, 1921, issued November 6, 1923. Assigned to Augustus Taylor, San Francisco, California.

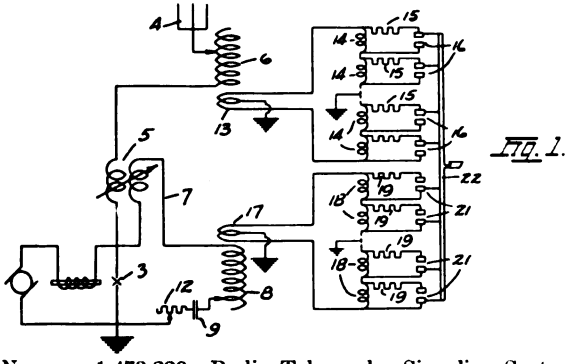

FIG. 1.

NUMBER 1,473,220—Radio Telegraphy Signaling System

RADIO TELEGRAPHY SIGNALING SYSTEM utilizing an arc at a radio transmitting station. The arc transmitter is provided with the usual oscillatory radiating circuit. An oscillatory non-radiating circuit is connected with the source of oscillations. A loop circuit is inductively coupled to one or both of the oscillatory circuits and a plurality of reactors, connected in series, are provided in each loop circuit. Shunting keys are arranged simultaneously to short-circuit each of the reactors for oppositely varying the impedances of the loop circuits, thus enabling the non-radiating circuit to absorb energy during the intervals between which the radiating circuits are transmitting energy.

1,473,417—F. G. Beetem, filed June 16, 1920, issued November 6, 1923.

RADIO RECEIVING APPARATUS, in which the filament current for the electron tubes in the receiver is maintained constant by a circuit connection which includes a resistor of high positive temperature coefficient connected between the source of energy and the filament circuit. A resistor is also connected in parallel with the filament to divert a portion of the current whereby the current in the filament is maintained constant at a different value from that of the current from the source.

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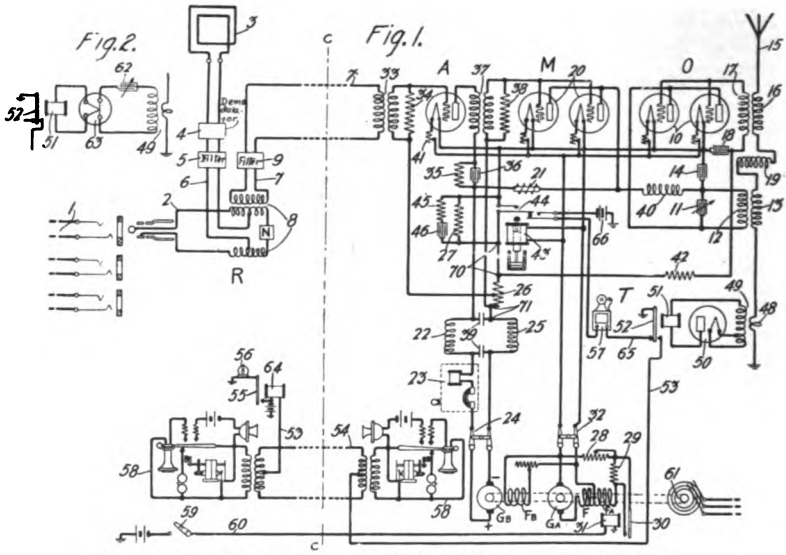
1,473,433—A. W. Kishpaugh, filed May 3, 1921, issued November 6, 1923, Assogned to Western Electric Company, Incorporated.



NUMBER 1,473,433—Carrier Wave Transmission System

CARRIER WAVE TRANSMISSION SYSTEM, in which the radio apparatus is located at a central station and arranged to be controlled from a switchboard which may be located at a distance. The generators which supply filament heating current and space current for the transmitting tube are controlled from the switchboard. An indicator is provided for showing the presence of the proper frequency in the transmitting antenna and for giving an alarm when the current varies or the frequency changes materially. Provision is also made for preventing injury in the transmitter circuits from excessive space current before the cathodes have become fully energized.

1,473,719—R. R. Beal, filed February 19, 1920, issued November 13, 1923. Assigned to Augustus Taylor of San Francisco, California.

RADIO TELEGRAPHY by means of an arc transmitter. This patent shows a circuit for an arc transmitter in which the continuous signaling waves are converted into wave trains or wave groups at the transmitter so that signals so transmitted are readily received and identified by all receiving stations. The

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signaling is accomplished by varying the core losses in a magnetic core having a winding associated therewith and connected in the antenna circuit. An audio frequency flux is impressed on the core from a keying circuit reducing the hysteresis and the effective resistance of the antenna circuit, thus increasing the antenna current to signaling value. Normally the losses effectively increase the resistance of the antenna circuit to such a value that current therein is reduced below signaling value.

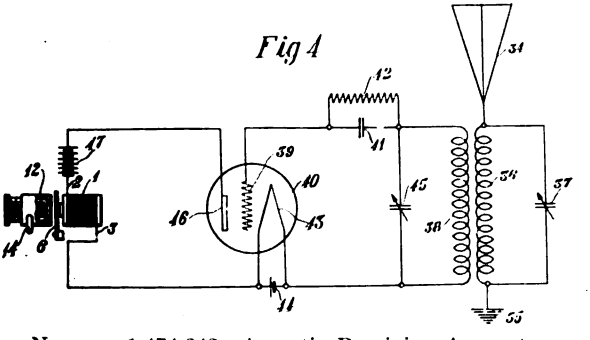
1,473,921—J. Bethenod, filed May 6, 1922, issued November 13' 1923.

HIGH FREQUENCY SIGNALING SYSTEM utilizing high frequency generators such as alternators for the transmission of signals. A synchronizing connection is provided, enabling a plurality of high frequency generators to be operated in parallel. The synchronizing connection consists of a pair of impedances in series shunted by a condenser and an inductance in series. The radiating system is connected to the junction point of the impedances.

Reissue 15,722—R. A. Heising, filed December 27, 1918, issued November 13, 1923. Assigned to Western Electric Company, Incorporated.

SYSTEM FOR PRODUCING MODULATED WAVES in an arc transmitter. An electron tube circuit is provided in association with an arc oscillator. An auxiliary arc oscillating circuit is also provided which has a period which tends to vary as the signal wave varies operating to control the main arc discharge for transmission of signals.

1,474,242—C. A. Culver, filed October 23, 1919, issued November 13, 1923.



NUMBER 1,474,242—Acoustic Receiving Apparatus

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Acoustic Receiving Apparatus, which consists in an electromagnet monotone system having an electromagnetically vibrated reed supported at one end in order to be fully responsive to vibrations of a single frequency. An electromagnet is mounted in a position to vibrate the reed from one side of the reed. A resonance chamber is located in a position on the other side of the reed and serves to amplify the intensity of sounds produced by the reed. The resonance chamber may be changed in period in order accurately to approach the period of the reed. Signals may be received by means of a listening tube connected with the resonance chamber.

1,474,293—F. Reynolds, filed November 10, 1921, issued November 13, 1923. Assigned one-half to the Silica Syndicate, Ltd., of London, England.

Thermionic Valve, in which provision is made for the expansion and contraction of the electrodes under conditions of temperature change. The electrodes within the tube are telescopically supported in members which are secured to the tube. An arrangement of springs within the telescopic members is provided whereby the electrodes are free to expand or contract as the temperature within the tube may vary.

1,474,382—H. J. Round, filed March 31, 1920, issued November 20, 1923. Assigned to Radio Corporation of America.

Apparatus for Radio Telegraphy and Telephony at the receiving station in which a coupled primary and secondary circuit is employed and a slotted metallic sheath positioned to surround the secondary winding. The metal sheath is connected with the filament circuit of the receiver. The metal sheath operates to eliminate the effect of short forced waves upon receiver.

1,474,486—B. MacPherson, filed June 3, 1919, issued November 20, 1923. Assigned to Wireless Specialty Apparatus Company.

Electrical Condenser of the stacked plate and sheet type, The stack is divided into sections, the sections being in series for high potential service. The invention resides in connecting the sections together by providing a metal sheet between adjacent sections, the sheets being connected at one end with the terminal of one section and at the opposite end with the terminal of the next succeeding section.

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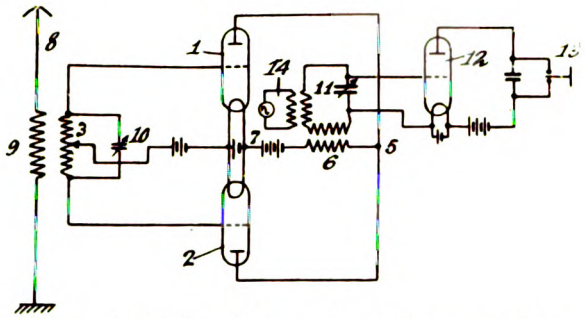
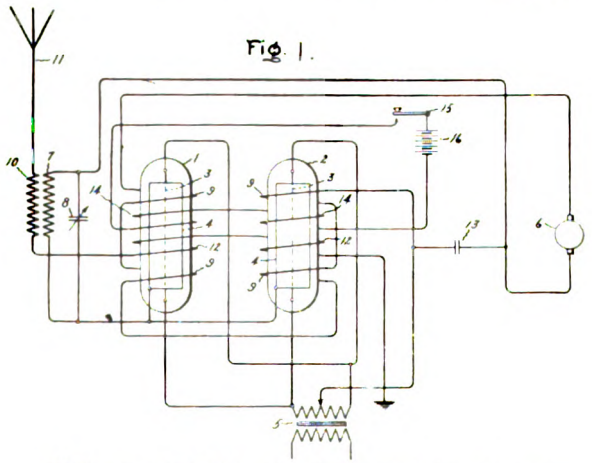
1,474,726—A. Meissner, filed August 8, 1922, issued November 20, 1923. Assigned to Gessellschaft, für Drahtlose Telegraphie m. b. h. of Berlin, Germany.



NUMBER 1,474,726—Method of and Arrangement for Receiving Electrical Gscillations

METHOD OF AND ARRANGEMENT FOR RECEIVING ELECTRICAL OSCILLATIONS at a radio station, wherein the number of channels of communication simultaneously operable side by side may be quadrupled over the number of channels available with the usual heterodyne reception. In the present system the signals are received and then increased in frequency, before the heterodyne, and a beat frequency derived from the combination of the increased frequency and a local generator which is detected for observing the signals. The channels are separated by differentially increasing and thereby separating the frequency of the signals in the receiving station.

1,475,164—W. R. G. Baker, filed June 9, 1922, issued November 27, 1923. Assigned to General Electric Company.



NUMBER 1,475,164—Signal Transmitting System

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SIGNAL TRANSMITTING SYSTEM, utilizing vacuum tubes for the generation of radio frequency currents in which an electron current is controlled by a magnetic field. The oscillatory circuit is connected with the electrodes of a magnetron, and a winding is provided surrounding the magnetron for producing a polarizing magnetic field. A third winding is provided surrounding the magnetron for producing a magnetizing field of sufficient value to interrupt the production of oscillations in accordance with telegraphic signals.

1,475,219—R. Bown, filed November 1, 1920, issued November 27, 1923. Assigned to American Telephone and Telegraph Company.

RADIO SIGNALING SYSTEM, arranged for multiplex operation with a plurality of transmission channels under centralized control and arranged to operate in co-operation with a corresponding number of secondary radio stations without interference. The transmitting frequency of a secondary station is controlled by transmitting from the primary station a modulated wave of fixed frequency. The received modulated wave is beaten at the secondary station with the unmodulated transmitted wave of the secondary station. The frequency of the unmodulated transmitted wave is adjusted until the difference in frequency between the received modulated wave and the unmodulated transmitted wave is of such modulation as to pass thru a selective circuit associated with the receiving circuit of the secondary station. In this manner interference between stations is prevented.

1,475,297—R. B. Goldschmidt, filed June 25, 1920, issued November 27, 1923.

RADIO SELECTION SYSTEM, which employs at the transmitting station a regulator which insures a uniform succession of trains of transmitted waves, and at the receiving station a regulator which causes the receiver to come into operation only during the time of transmission of such trains. The system is intended to eliminate interference from extraneous signals.

1,475,448—Harold Rowntree, filed May 16, 1921, issued November 27, 1923.

ELECTRICAL TRANSMISSION OF COMMUNICATIONS, consisting of a method in which a plurality of characteristics of a transmitted carrier current are continuously altered in a predetermined

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sequence of nature, type or degree of alteration. The carrier current is simultaneously modulated in accordance with the signal to be transmitted. The object of the invention is to secure greater selectivity for providing more channels of communication.

1,475,583—C. A. Hoxie, filed May 20, 1921, issued November 27, 1923. Assigned to General Electric Company.

VARIABLE CURRENT GENERATOR, wherein the amount of current or voltage produced at any instant is controlled by a screen placed in the path of light travelling from a suitable source to a photo-electric cell. This screen is provided with specially formed cut-away or transparent portions and is arranged to be moved so that the light which passes thru these portions and falls upon the photo-electric cell is caused to vary in a manner corresponding to the variation which it is desired to produce in the current or voltage. The photo-electric cell is connected to control an electron tube circuit.

1,475,632—H. B. Herty, filed February 10, 1922, issued November 27, 1923.

PROTECTIVE DEVICE FOR RADIO RECEIVING SYSTEMS, in which a thermal device is connected in the antenna system to control a relay circuit which operates on excess current for short-circuiting the element which couples the receiver to the antenna, thereby avoiding injury which might result to the receiver in case of excess energy in the antenna circuit.

1,476,003—De Loss K. Martin, filed August 18, 1922, issued December 4, 1923. Assigned to American Telephone and Telegraph Company.

RADIO SIGNALING CALL SYSTEM, in which any one of a number of remote stations may be called from a central radio station. The system includes a plurality of stations arranged for intercommunication, with calling means for establishing channels between one station and any other which comprises a given combination of voice frequencies on a plurality of different carrier frequencies.

1,476,156—H. P. Donle, filed April 7, 1921, issued December 4, 1923. Assigned to Connecticut Telephone and Electric Company, Incorporated.

RADIO FREQUENCY DEVICE which consists of a vacuum tube containing only a cathode within the evacuated space. The anode

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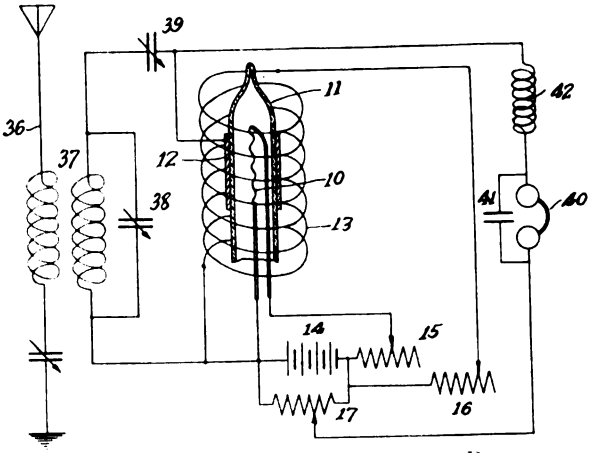
is placed in intimate contact with the outer surface of the tube and connected with the cathode. An adjustable magnetizing coil is provided which surrounds the anode outside of the tube.

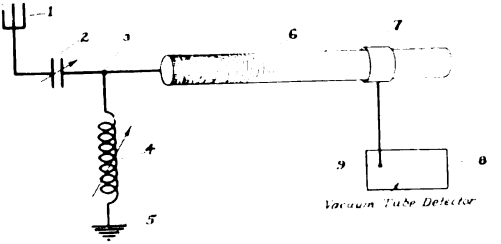


NUMBER 1,476,156—RADIO FREQUENCY DEVICE

1,475,027—E. F. Randall, filed June 7, 1923, issued November 20, 1923. Assigned to American Radio and Research Corporation.

DETECTOR FOR RADIO SIGNALS of the crystal type in which the connecting member held in contact with the rectifying crystal is maintained in position by a permanent magnet. The conducting contact member may be moved in different positions and held in such position by the magnetic attraction between the contact member and the permanent magnet.

1,476,691—L. Cohen and J. O. Mauborgne, filed January 26, 1921, issued December 11, 1923.



NUMBER 1,476,691—ELECTRICAL SIGNALING

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ELECTRICAL SIGNALING, in which a "wave coil" is employed for eliminating static disturbances at the receiver. The receiving circuit may include an antenna tuned by a series condenser, a loop circuit, and a ground connection, constituting the antenna circuit. The wave coil is connected to a point on the antenna circuit with the receiving apparatus associated with the wave coil. The operation is as follows: The antenna circuit 1, 2, 3, 4, and 5, is tuned to the frequency of the signal which it is desired receive and if the capacity of the condenser 2 is small the inductance 4 large, the resonance potential across the condenser 2 may be made relatively very large, that is, many times the voltage induced in the antenna by the electromagnetic waves of the signal. Connecting the wave coil 6 to the point 3 on the antenna circuit, a high potential is impressed on the wave coil and part of the signal energy is transmitted over the wave coil effecting a wave development on the coil. Since the signal energy is transmitted at a high potential, the current in the wave coil is correspondingly small and the efficiency of transmission is accordingly large, comparatively little energy is lost in the transmission. By suitably adjusting the length of the coil in relation to wave length of the signals, one or more maximum potential points will occur on the coil. While good results may be obtained if the length of the wave coil is less than one-quarter of the wave length, it is preferable to have the coil of sufficient length to secure the development of at least a quarter wave length. By connecting the detector 8 to a maximum potential point on the wave coil, which is accomplished by sliding the ring 7 along the coil, a further increase in the receiving efficiency of the system is thus obtained. In using a three-electrode vacuum tube detector, the grid 9 is connected to the ring. When any other electrical disturbances act on the antenna, such as interfering signals or electrostatic disturbances, the effect on the detector is small, for the reason that the antenna circuit is not in resonance for the interference effects and hence the potential across the condenser 2 produced by these effects is relatively small. The effect is more in the nature of a rush of current flow thru the antenna circuit, part of which will necessarily be transmitted over the wave coil, but because of the large resistance of the wave coil the energy is quickly dissipated and the disturbance is very largely attenuated before it reaches the part of the wave coil which is connected to the detector, and therefore the effect on the detector is small.

1,477,316—F. Conrad, filed July 11, 1922, issued December 11, 1923. Assigned to Westinghouse Electric and Manufacturing Company.

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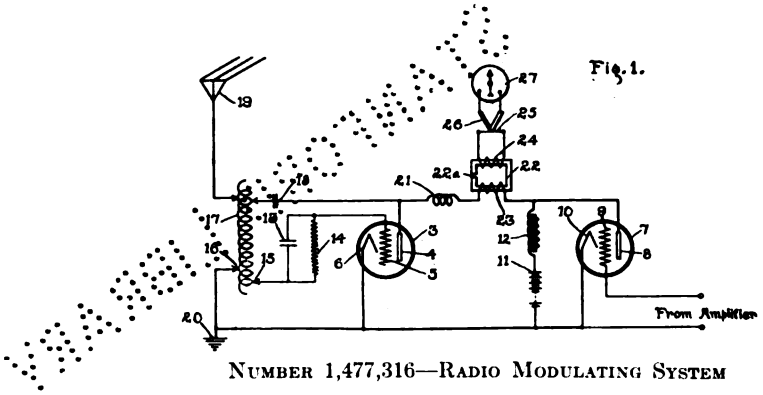
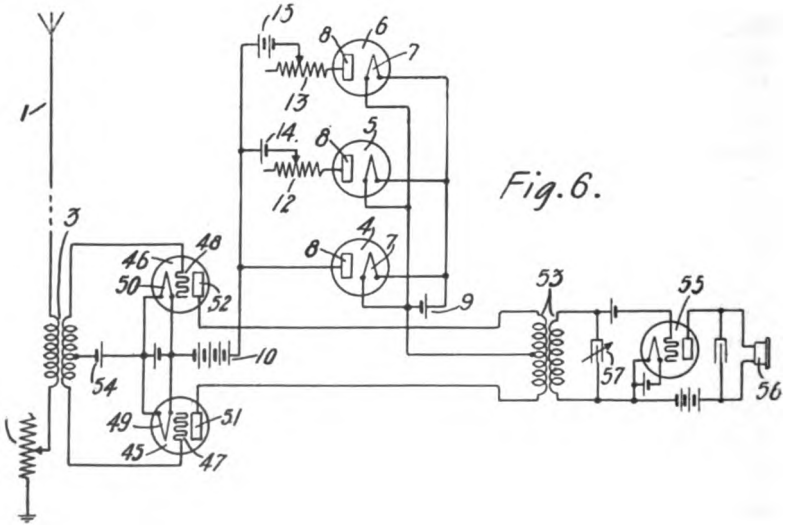


NUMBER 1,477,316—RADIO MODULATING SYSTEM

RADIO MODULATION SYSTEM, in which a meter is provided for registering the alternating current component of the modulated current supplied to the oscillator tube at the transmitter. The meter is calibrated to indicate directly the ratio of the effective alternating current component to 0.7 times the steady direct current component. A current transformer is connected in circuit with the oscillator and may operate an oscillograph or meter to indicate the ratio of the actual modulated current to the minimum modulated current possible without undesirable distortion. The apparatus is described in connection with a broadcasting station where the characteristics of the transmitter are always observable by the operator in charge of the station controls.

1,476,721—D. Loss K. Martin, filed November 23, 1921, issued December 11, 1923. Assigned to American Telephone and Telegraph Company.



NUMBER 1,476,721—FREQUENCY CONTROL SYSTEM

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FREQUENCY CONTROL SYSTEM for producing synchronism between a locally generated frequency and a controlling frequency transmitted from a distant point, which consist in adjusting an oscillator so that it will normally produce a frequency in the neighborhood of the controlling frequency, transmitting the controlling frequency thru a one-way device, and in impressing the controlling frequency thus passed upon the oscillator whereby the oscillator will oscillate in synchronism with the controlling frequency.

1,477,314—E. F. W. Alexanderson, filed June 24, 1921, issued December 11, 1923. Assigned to General Electric Company of New York.

RADIO RECEIVING SYSTEM having a circuit arrangement for preventing interference by signals from an undesired transmitting station. A multiple tuned circuit is connected to the antenna which is resonant to the frequency of the disturbing effect to be eliminated. A series tuned circuit is also connected to the antenna and it is also resonant to the frequency of the disturbing effect to be eliminated. The electromotive forces derived from both of the resonant circuits are impressed upon the same receiving circuit. By properly proportioning the circuits the voltage drops produced across these circuits by the desired signal may be made to be substantially opposite in phase, so that they will add in the receiving circuit. The magnitude of the voltage drop across the multiple tuned circuit will also be less than the voltage drop across the series tuned circuit. The signal which has a frequency only slightly different from that of the interfering wave may be received with practically full intensity at the same time that interfering wave is neutralized.

1,477,645—R. E. Hall, filed August 13, 1919, issued December 18, 1923. Assigned to Hall Research Corporation of Delaware.

SIGNAL RECEIVING SYSTEM AND METHOD, in which a plurality of independent receiving stations located at separate points are all tuned to the same transmitting station. The separate relays operate to synchronously respond to the received signal and operate a central recorder connected in a relay circuit with all of the receivers. Static disturbances which are local to one of the stations may not exist at one of the other stations. Such disturbances being unsynchronized will not affect the central recorder, while a desired signaling frequency will synchronously

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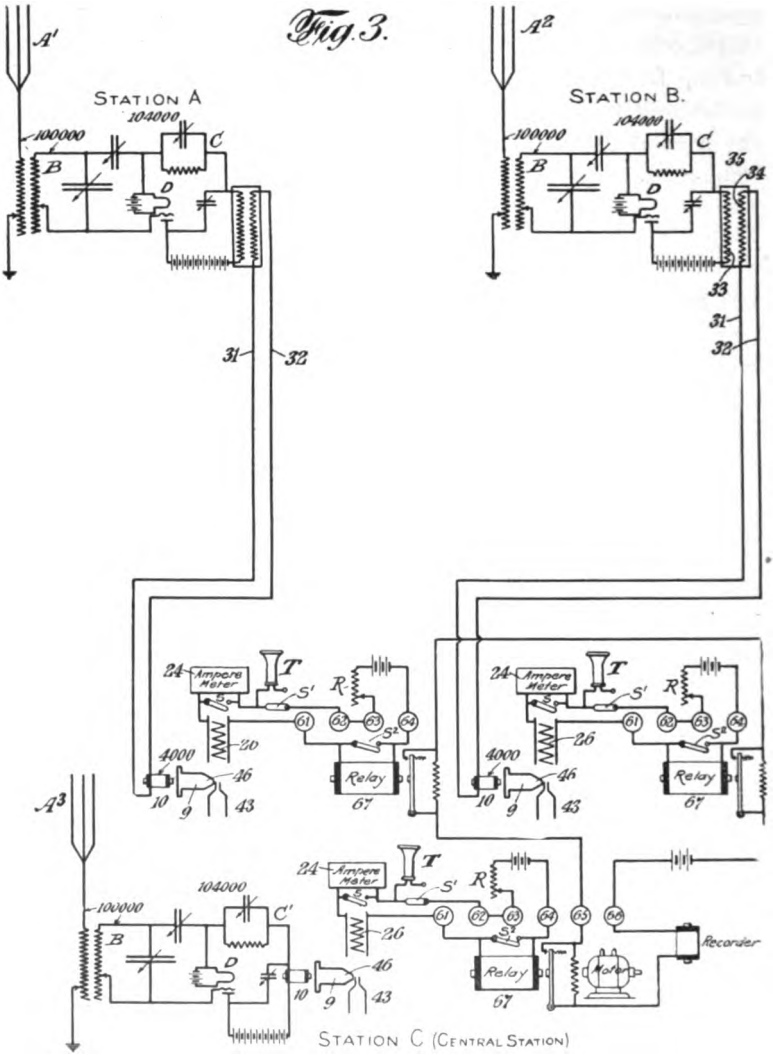
operate the several receivers and thereby control and operate the central recorder.



NUMBER 1,477,645—SIGNALING RECEIVING SYSTEM AND METHOD

1,477,826—R. J. Heitzman, filed June 29, 1922, issued December 18, 1923.

DETECTOR STAND of the crystal type in which a lever is pivoted upon a vertical standard and one end of the lever adjusted up and down by rotation of a cam against the action of a

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spring. A contact wire is attached to the opposite end of the lever and is capable of variable adjustment against the rectifying crystal.

1,477,868—H. P. Donle, filed May 9, 1919, issued December 18, 1923. Assigned to Connecticut Telephone and Electric Company.

METHOD AND APPARATUS FOR INCREASING ELECTRONIC EMISSION from an electrode within the glass bulb of a thermionic valve which consists in extracting a metallic element from the material of the glass bulb and depositing the same upon the cathode. The patent states that with this arrangement the electronic emissivity is greatly increased.

1,477,869—H P. Donle, filed February 3, 1923, issued December 18, 1923. Assigned to Connecticut Telephone and Electric Company, Incorporated.

ELECTRONIC EMISSION secured in increased proportions by attracting to the cathode of the tube electrically charged particles of sodium. The object of the invention is to provide metallic ions within the tube and thus, in addition to furnishing a basis for ionization effects, increase the electron emission of the cathode for a given temperature.

1,477,899—C. W. Rice, filed July 27, 1921, issued December 18, 1923. Assigned to General Electric Company.

HIGH FREQUENCY SIGNALING SYSTEM, in which the stray or static ratio is materially improved at the receiver. A circuit is provided at the receiver which includes an artificial transmission line having an effective length equal to a plurality of wave lengths of the signals to be received. The received signals are impressed on this transmission line and then the signaling currents derived from more than two points in the transmission line impressed upon the receiving apparatus.

1,478,029—Lee de Forest, filed July 2, 1920, issued December 18, 1923.

RADIO RECEIVING SYSTEM for receiving undamped wave trains in the form of a musical note without the intermediary of a tikker, tone wheel, or mechanical interrupter of any kind, and further, without using the principle of heterodyne or beat note reception. The patent describes a circuit in which the received

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radio frequency current is utilized to interrupt itself, thereby to produce current impulses of audio frequency.

1,478,047—J. Mills, filed December 22, 1920, issued December 18, 1923. Assigned to Western Electric Company, Incorporated.

Radio Receiving System for avoiding interference due to static or other atmoshperic conditions. A plurality of sets of incoming waves are separately received at different stations and caused to modulate separate carrier waves which are transmitted to the same receiving station, where they are combined to secure the signal with suppression of interference.

1,478,050—E. L. Nelson, filed April 26, 1922, issued December 18, 1923. Assigned to Western Electric Company, Incorporated.

Modulation Circuits and Measurements, in which a device is employed at the radio transmitter which has an indicator moving over a calibrated scale. This device is responsive to current intensities and is included in the modulation circuit to indicate the degree of modulation. The patent points out that this measuring device is to be preferred over the oscillograph for studying output at the broadcasting station.

1,478,072—H. J. Van Der Bijl, filed March 20, 1918, issued December 18, 1923. Assigned to Western Electric Company, Incorporated.

Vacuum Tube construction in which the grid and plate are mechanically reinforced against distortion from their original shape by means of ridges on the plate and grid. These reinforcing means serve to always maintain the electrodes in the same spaced relation and at the same time they offer substantially no obstruction to the passage of electrons.

1,478,076—H. W. Weinhart, filed August 7, 1919, issued December 18, 1923. Assigned to Western Electric Company, Incorporated.

Electron Discharge Device, in which an inwardly projecting portion is provided within the tube upon which the electrodes are supported by means of bands which are independently clamped to the inwardly projecting portion.

1,478,087—W. Wilson, filed November 21, 1919, issued December 18, 1923. Assigned to Western Electric Company, Incorporated.

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VACUUM TUBE construction, in which the cathode is formed by a plurality of intertwined twisted ribbons of thermionically active material. The object of the invention is to provide a cathode which has the property of large electron emissivity at a minimum temperature.

1,478,342—R. C. Lewis, filed January 31, 1921, issued December 18, 1923. Assigned to Coto-Coil Company of Boston, Massachusetts.

ELECTRICAL AIR CONDENSER of the variable type in which the plates are supported in a single unitary mounting, consisting of a spider-member having an extended bearing hub at the center and three arms extending therefrom. The stationary plates are supported from the extremities of the arms and the movable plates are journalled in the bearing at the center of the spider.

LIST OF RADIO TRADE MARKS PUBLISHED BY PATENT OFFICE PRIOR TO REGISTRATION

(The numbers given are serial numbers of pending applications)

185,394—"RADIODYNE" in ornamental design for radio receiving sets. Western Coil and Electrical Co., Racine, Wisconsin. Claims use since August 19, 1923. Published November 13, 1923.

185,822—"GIANT" for crystal detectors. Foote Mineral Co., Philadelphia, Pennsylvania. Claims use since May 11, 1923. Published November 13, 1923.

185,823—"VARIO-TENSER" in ornamental design for cat whisker for detector. Foote Mineral Co., Philadelphia, Pennsylvania. Claims use since May 11, 1923. Published November 13, 1923.

184,252—"MARLE" in ornamental design for transformers. Marle Engineering Co., Orange, New Jersey. Claims use since May 19, 1922. Published November 13, 1923.

184,330—"MICRO-TUNE" for variable condensers. The Fett and Kimmel Co., Bluffton, Ohio. Claims use since May 14, 1923. Published November 13, 1923.

176,781—"RADIO ARGENTITE," a crystalline compound for detectors. J. Byrne Curry, doing business as Curry and

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Coutellier, Los Angeles, California. Claims use since July 1, 1922. Published November 27, 1923. (Not subject to opposition.)

165,987—"DICTOGRAPH RADIO" in ornamental design for telephonic headsets. Dictograph Products Corporation, New York, New York. Claims use since May 1, 1922. Published November 27, 1923.

185,665—"MURAD" for radio receiving apparatus. Mu-Rad Laboratories, Inc., Asbury Park, New Jersey. Claims use since November 20, 1921. Published November 27, 1923.

186,856—"R. C. Co." for electrical condensers. Radio Condenser Company, Camden, New Jersey. Claims use since February 1, 1923. Published November 27, 1923.

187,109—"L-ANCO" in ornamental design for vario-couplers. Lanco Coupler Co., Lancaster, Pennsylvania. Claims use since October 15, 1923. Published November 27, 1923.

186,974—"BRADLEYLEAK" for grid leaks. Allen-Bradley Company, Milwaukee, Wisconsin. Claims use since September 1, 1923. Published November 27, 1923.

179,912—"FOOTE" for radio crystals. Foote Mineral Co., Philadelphia, Pennsylvania. Claims use since May 1, 1922. Published November 27, 1923. (Not subject to opposition.)

168,936—"THE NORTH AMERICAN RADIO & SUPPLY CORPORATION" in ornamental design for Radio apparatus. The North American Radio and Supply Co., New York, Claims use since June 21, 1922. Published December 18. 1923.

181,519—"EKKO" in ornamental design for radio phonograph apparatus. The Ekko Company, Chicago, Illinois. Claims use since July 1, 1922. Published December 18, 1923.

183,040—Triangular design for dry batteries. National Carbon Company, New York. Clains use since June 1, 1923. Published December 18, 1923.

184,108—"BASUB" in ornamental design for apparatus for adapting commercial alternating current for use in radio receiving circuits as a substitute for a storage battery.

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Simplex Electrical Laboratories, New York and Brooklyn. Claims use since on or about July 1, 1923. Published December 18, 1923.

185,358—"CELERUNDUM—THE CRYSTAL WITH A SOUL" for radio detectors. The Ferris Radio Research Laboratory, Boston, Massachusetts. Claims use since July 1, 1923. Published December 18, 1923.

186,024—"EVEREADY-THREE" for dry batteries. National Carbon Company, Inc., New York, New Qork. Claims use since June 1, 1923. Published December 18, 1923.

177,606—"TRUTONE" in ornamental design for radio loud-speaking horns. Claims use since July 5, 1922. Published December 18, 1923. Sadler Mfg. Co., San Francisco, California. (This mark is nor subject to opposition.)

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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 12 APRIL, 1924 Number 2

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RECEIVING MEASUREMENTS AND ATMOSPHERIC DISTURBANCES AT THE BUREAU OF STANDARDS, WASHINGTON, SEPTEMBER AND OCTOBER, 1923*

By

L. W. Austin

(Chief, Radio Physical Laboratory, Bureau of Standards, Washington, District of Columbia.)

(Communication from the International Union of Scientific Radio Telegraphy)

Changes are reported to have been made at Nauen resulting in an increase of radiation height to 170 m. (557 feet), while the current has been reduced to 270 amperes. This gives a net decrease in calculated field intensity, in America, of about twenty-five percent. The wave frequency is now 23.4 kc. ($\lambda = 12{,}800$ m.).

The main points to be noted in the September and October measurements are:

(a) The continued increase in strength of Lafayette, already mentioned, which is more than twice as strong in the forenoon (all daylight path) as in the same months last year, with no increase in antenna current.

(b) The unchanged strength of Nauen, notwithstanding its decrease in antenna current.

(c) That the afternoon fading of Lafayette is practically the same as last year, while that of Nauen continues nearly a month later; and that the heavy atmospheric disturbances have also continued much later, altho they were less severe during the summer.

*Received by the Editor, January 15, 1924.

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FIELD INTENSITY OF NAUEN AND OF DISTURBANCES ($f = 23.4$ kc., $\lambda = 12.800$ m.) IN SEPTEMBER, 1923, IN MICROVOLTS PER METER

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
1	82	60	....	....
4	29	45	*	120
5	64	80	*	450
6	**	60	*	200
7	42.7	45	*	200
8	60	65	....	....
10	60	40	21.5	250
11	72.7	25	*	200
12	64	30	*	280
13	55.7	30	*	200
14	55.7	25	*	160
15	51.2	30	....	....
17	38.2	20	*	165
18	24.3	15	8.0	80
19	38.2	40	20.4	200
20	28	35	2.0	150
21	29	80	4.5	350
22	24.3	20	12.2	200
24	29.5	50	2.0	300
25	51.5	20	26.0	180
26	51.5	40	17.0	130
27	25.7	45	23.5	100
28	42.8	30	9.9	130
29	42.8	10	19.2	95
Average	46.2	39.1	7.9	197

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FIELD INTENSITY OF LAFAYETTE AND OF DISTURBANCES ($f = 15.9$ kc., $\lambda = 18{,}900$ m.) IN SEPTEMBER, 1923, IN MICROVOLTS PER METER

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
1	198	60	....	....
4	120	48	114	100
5	145	85	24.5	500
6	**	60	96	180
7	180	48	54.2	250
8	**	80	....	....
10	168	40	96	200
11	235	20	90	180
12	120	36	51.5	300
13	120	25	72.2	200
14	132	30	102	160
15	120	40	....	....
17	**	25	108	165
18	**	20	140	80
19	213	50	153	220
20	188	40	38.0	150
21	160	90	140	400
22	102	25	72	250
24	207	60	126	360
25	190	28	120	200
26	126	50	162	155
27	114	90	120	130
28	148	35	165	130
29	196	15	165	115
Average	159	45.8	105.2	210.7

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Field Intensity of Nauen and of Disturbances ($f = 23.4$ kc., $\lambda = 12{,}800$ m.) in October, 1923, in Microvolts per Meter

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
1	51.5	50	26.5	200
2	42.8	20	5.4	130
3	42.8	45	17.0	95
4	51.5	25	27.0	85
5	55.7	20	**	110
6	22.8	35	24.0	....
8	26.7	15	55.7	30
9	55.7	25	42.7	45
10	51.5	30	25.0	60
11	51.5	15	19.0	60
12	60.0	20	16.7	180
13	51.5	15	18.0	65
15	26.7	35	13.6	100
16	17.0	65	2.0	180
17	34.0	15	30.0	30
18	**	15	17.0	35
19	34.2	10	30.0	25
20	35.7	15	30.0	30
22	21.4	15	17.0	40
23	25.7	20	32.0	65
24	42.8	10	17.0	55
25	31.0	10	7.0	100
26	28.0	10	5.0	65
27	25.7	15	17.0	45
29	33.0	20	9.3	100
30	30.0	35	**	55
31	33.2	20	18.0	45
Average	37.4	23.1	20.8	78.1

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FIELD INTENSITY OF LAFAYETTE AND OF DISTURBANCES (f=15.9 kc., λ=18,900 m.) IN OCTOBER, 1923, IN MICROVOLTS PER METER

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
1	196	60	160	250
2	204	35	120	160
3	212	50	120	110
4	220	30	132	100
5	220	25	102	150
6	165	40	168	....
8	165	20	147	35
9	170	30	140	65
10	195	30	96	70
11	126	20	84	80
12	186	25	114	210
13	160	20	132	65
15	90	45	102	150
16	153	80	132	200
17	212	20	150	35
18	170	15	180	45
19	140	15	174	30
20	114	20	138	45
20	90	15	200	50
23	120	25	151	80
24	**	12	66	60
25	**	15	39	150
26	198	15	152	65
27	**	20	102	60
29	160	30	120	100
30	**	40	102	65
31	**	25	150	45
Average	166.6	28.8	128.5	95.2

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RATIO OF AVERAGES

Stations	Signal P. M. / A. M.	Disturbance P. M. / A. M.	A. M. Signal / Dist.	P. M. Signal / Dist.
		September		
Nauen	0.17	5.1	1.2	0.04
Lafayette	0.66	4.6	3.5	0.50
		October		
Nauen	0.55	3.4	1.6	0.27
Lafayette	0.77	3.3	5.8	1.35

SUMMARY: The signal field strengths and the corresponding strengths of atmospheric disturbances are given for the Nauen and Lafayette stations for September and October, 1923.

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SHORT PERIOD VARIATIONS IN RADIO RECEPTION*

By

Greenleaf W. Pickard

(Consulting Engineer, The Wireless Specialty Apparatus Company Boston, Massachusetts)

(Communication from the International Union of Scientific Radio Telegraphy)

In any continuous measurement of the field intensity from a distant transmitter, we find both periodic and irregular variations, often of great amplitude, and always of complex form. Some of these changes in the transmission coefficient of the radio circuit may be definitely correlated with solar and terrestrial happenings; for example, transmission is in general better in winter than in summer, better at night than during the day, and is usually adversely affected at the time of intense aurora. These fluctuations are less prominent at the lower transmission frequencies, and increase until several thousand kilocycles is reached. Then, altho the data at our disposal today is quite limited, they appear to decrease, until at ten thousand kilocycles, more or less, the transmission becomes quite uniform.[1] Altho there is little doubt but that these fluctuations are due to changes occurring somewhere in our atmosphere, we are to-day in ignorance as to the mechanism involved.

If we confine our attention to frequencies between five hundred and fifteen hundred kilocycles, that is, to the band now principally filled with radiophone broadcasting, and to overland transmission at distances between one hundred and one thousand kilometers (62 to 620 miles), we find that the average midwinter field intensity is about five times greater than in midsummer.

In the daytime, altho transmission is relatively free from the large amplitude short period variations commonly known as "fading," "swinging" or "soaring," there is usually a slow change from hour to hour, which in the majority of cases appears as a

*Received by the Editor, November 8, 1923. Presented before The Institute of Radio Engineers, New York, December 12, 1923.

[1] "Radio Telegraphy," Marconi, Proceedings of The Institute of Radio Engineers, volume 10, August, 1922, pages 235 to 236.

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gradual decrease from morning to night, the morning intensities often being twice those of the afternoon.

About half an hour before sunset at the receiving point, the weak and relatively constant field from the distant station begins to show marked short period fluctuations which grow in amplitude from minute to minute until, soon after sunset, they usually assume grotesquely large amplitudes. The principal change from day to night is an increase in field intensity; in general, the lower limit of the night-time field is approximately the same as the late afternoon field, altho from time to time there will be found a momentary fall to a much lower value than at any time during the day. The upper limit of the night-time field is not so definite; it may be ten, a hundred or even on occasion thousands of times greater than the daytime intensity, depending upon the distance and the character of night.

The amplitude of the short period variations, that is, those fluctuations ranging in duration from seconds to tens of minutes, is principally controlled by the distance between the transmitter and receiver, and this is true of both night and day transmission. At less than eleven kilometers (6.9 miles) from a broadcasting station, there is a well-defined short period fluctuation in intensity during night-time transmission, which is practically absent during the day, and which on some evenings shows an amplitude of ten percent or over. As the distance increases, the amplitude of the variations also increases, becoming readily observable by day at distances of fifty kilometers (33 miles), more or less.

At first there is no change in the character of the fluctuation other than in amplitude, but when a distance of between one and two hundred kilometers (63 and 125 miles) is exceeded, the oscillations of periods ranging from seconds to a minute or two become less prominent, and the longer swings of minutes to tens of minutes are accentuated. At an ill-defined distance of perhaps one hundred and fifty kilometers (94 miles) the amplitude of the shorter period variations appears to be at a maximum.

When continuous records of night-time field intensity from a distant station are made at separated receiving points, I have found that altho the records are of the same general character they are not identical unless the receiving points are practically coincident. A separation of less than six hundred meters (1,970 feet) between two receiving points is sufficient to obliterate most of the detail resemblance between their records, save only for the more prominent long period swings. Equally, simultaneous records at a single receiving point from separated

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transmitters show few detail resemblances other than the seasonal and diurnal changes, altho here there is, of course, a difference in frequency.

A comparison of records made on different evenings shows that these evenings may be roughly classified as quiet, moderately disturbed, or highly disturbed for all stations within a two or three hundred kilometer (125 or 188 mile) radius. Whenever a station eleven kilometers (6.9 miles) distant shows marked short period fluctuations, more distant stations will also exhibit the same disturbances. On a quiet night the local station shows little or no variation in field intensity, and the more distant stations show principally long period disturbances.

The records of broadcast transmission which form the principal part of this paper have all been made by reception on non-directional open antennas, as my object here is to present only true changes in intensity of electric field, as distinct from the changes in wave-front which often accompany them. In a subsequent paper I intend to present records made on both stationary and rotating directional aerials, in which I have found striking changes in wave-front over very short intervals, some of these changes being very definitely correlated with changes in true field intensity.

The received current in the open antenna is impressed upon a crystal detector, directly if the field intensity is sufficient, but usually after from one to four stages of radio frequency amplification, and the rectified current from the detector is then passed thru a galvanometer. In one form of recorder employed for making these records, the registration is photographic, a beam of light reflected from the galvanometer mirror playing across a sheet of photographic paper wound on a slowly rotating drum, while in another and more portable form of recorder a pointer galvanometer is used, the registration being semi-manual.

One of the arrangements employed for recording is shown in Figure 1. The open antenna O includes a filter F, for minimizing interference from local stations, a series tuning condenser C 1, a loading inductance L 1, and a coupling coil L 2. A grounded shield S electrostatically separates the antenna circuit from the tuned grid circuit L 3, C 2 of the radio frequency amplifier train. The amplifier circuit here shown is that of Mr. Chester W. Rice,[2] which I have found very stable and constant, and in every way suitable for this rather exacting work. The voltmeters VM 1 and VM 2, which measure the filament and plate voltages in the

[2] United States patent, number 1,334,118.

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amplifier train, are essential adjuncts, as it is of first importance that the voltage amplification be kept rigidly constant over long periods.

The amplified radio frequency current resulting from this train is impressed upon a crystal detector *D*, and the rectified current from this detector then passes thru a reflecting galvanometer *RG*, provided with a constant impedance shunt *CIS* for adjusting the deflection to the width of the record sheet. Monitoring telephones *MP* are shunted across the galvanometer circuit, with a series condenser *C* 8 to prevent any drain of direct current from the galvanometer circuit. A light projector *P* 1, consisting of a six-volt 2-candle-power lamp, a pinhole diafram, and a lens, throws a beam of light upon the galvanometer mirror, from whence it is reflected to a sheet of photographic paper wound on a kymograph drum *K*, which is normally rotated once an hour by a clockwork motor *M*. A similar light projector *P* 2, provided

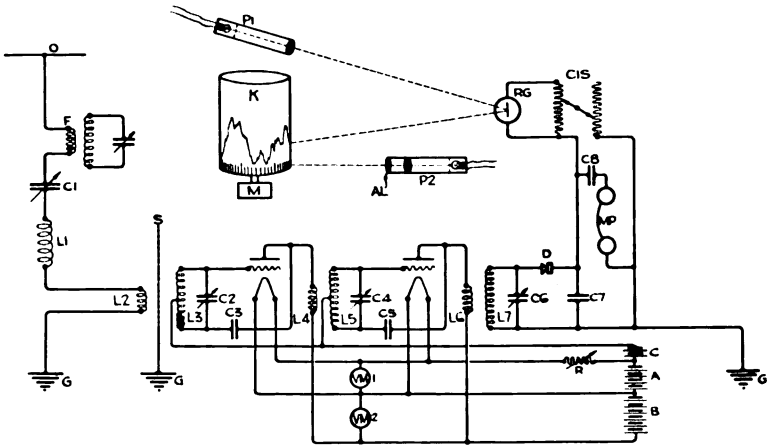

FIGURE 1

in addition with an astigmatic lens *A L*, flashes a short line of light once every minute on the lower edge of the sensitive paper, thus giving a time scale. The galvanometer employed is of fair sensitiveness, giving approximately one millimeter deflection at one meter for a current of 10^{-9} ampere, so that it is not necessary to overload the crystal detector, and so depart from the current-square relation between input and output. The kymograph, galvanometer, and light projectors are, of course, installed in a dark room, with shielded leads running in from the detector, which with its associated amplifier, tuner, and filter is placed in

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another room. A crystal detector preceded by a radio frequency amplifier train is extraordinarily stable in adjustment, and the one employed at my receiving station in Newton Centre, Massachusetts,* has operated without readjustment for over five months at a time, with no measurable change in sensitiveness.

Another form of recorder which has been extensively employed in this work, and which has the advantage over the photographic form of portability, freedom from photographic process and delay, and great simplicity, is due to Mr. H. S. Shaw, Jr., who has been associated with me in this work almost from its beginning. A photograph of this recorder is shown in Figure 2. A strip of paper, 9 cm. (3.55 in.) wide, passes around a drum

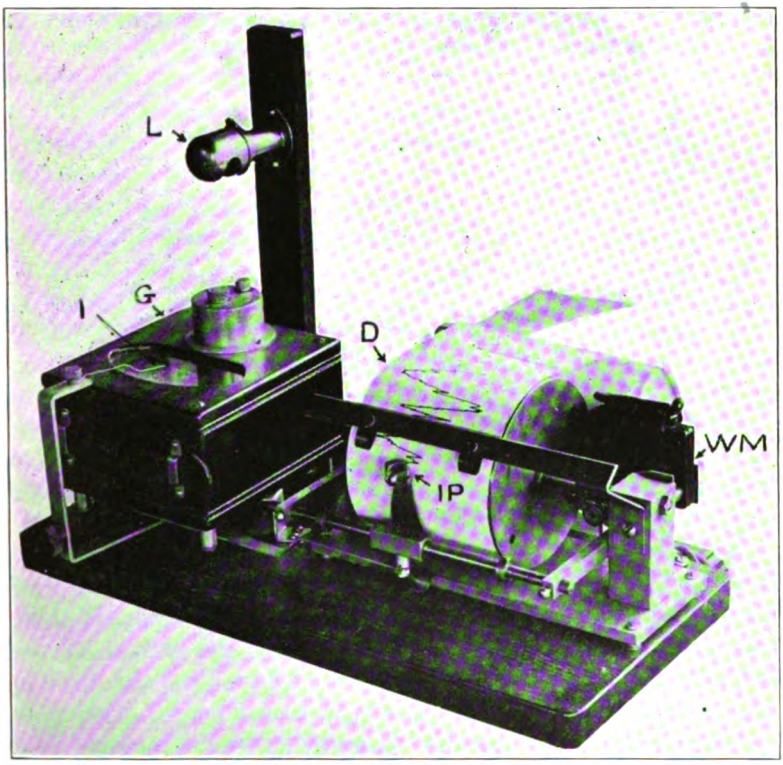

FIGURE 2

D, rotated by a worm and gear drive from a Warren Clock Motor *W M*. A pointer galvanometer *G* to the left of the drum is connected to the crystal detector, and movements of its pointer are followed manually by the index *I* (the shadow of which is

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cast on the galvanometer scale by the light L) which in turn transmits the motion to an inking pen IP, traversing the paper strip. It might be thought that this form of recorder would be rather tedious to operate, but after a few minutes' recording the movements become almost automatic, and a record an hour long is not at all tiresome. The time scale of this manual recorder is made more open than the photographic, being 1.6 cm. (0.63 in.) per minute, while the kymograph scale is only 0.5 cm. (0.19 in.) per minute.

With a fair intensity of electric field, that is, of the order of a thousand microvolts per meter, and a fair sized open antenna, excellent records may be made without any amplification at all, direct from the crystal alone, as may be seen from Figure 3, recorded by Mr. Shaw at Scituate, Massachusetts, from Schenectady, New York, 265 kilometers (165 miles) away.

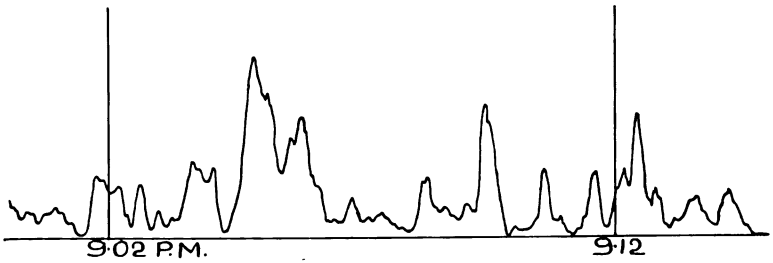

FIGURE 3

Another almost equally simple arrangement is to shunt the crystal detector across the plate impedance of a regenerative tube, and with this good records may be made with electric fields of a few hundred microvolts per meter.

As mentioned above, the ordinates of these records are very closely proportional to the square of the electric field, so that they might properly be termed energy records. For many of these records I have determined approximately the values of the ordinates in microvolts per meter, and I have used this data in determining the seasonal variation in transmission. Owing to the extraordinary complexity of many of these records, it is difficult to determine the mean value over any extended period by mere inspection, and the method which I have adopted is to measure the area of the space between the curve and its base or zero line with a polar planimeter, and divide the result by the length of the base line.

*Newton Centre is 11 kilometers (7 miles) in a westerly direction from Boston, Massachusetts.

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Altho the methods outlined above appear and are actually simple, it is not altogether easy to obtain an unbroken series of good records from distant stations. When the field from the distant station is weak, high amplification is necessary, and the outfit is then at the mercy of disturbances, such as local spark station, nearby amateur working, and even a violently oscillating receiver in the immediate neighborhood. Also, the record may start at a time of low field intensity, and the galvanometer shunt and amplification may be so pessimistically chosen that when the intensity rises most of the record will be "off scale."

Before taking up the records in detail, I wish to make clear the relation between the intensities shown on my records and those observed by ear, either from phones on the head or from a loud speaker. In general, the human ear is unable to discriminate between sounds of the same character, but of different intensities, unless the difference is quite large. For example, trained observers making comparisons by ear of telephone transmission over different circuits are usually unable definitely to determine differences of less than one mile of standard cable, that is, a change in amplitude of about ten percent. Such a comparison is, of course, made under the most favorable conditions possible, that is, the two transmissions are rapidly alternated by switching from one circuit to the other. Where there is no standard of comparison, and the observer is listening to a single transmission which is slowly varying in intensity, the change to be observable varies from about thirty to several hundred percent, depending principally on the rate of change and the absolute intensity of the sound in the ear. If the intensity is fluctuating rapidly, that is, every few seconds, discrimination is, of course, at its best, whereas if the change is a smooth one over a period of minutes, it will pass undetected unless quite large. Also, if the sound is very loud, discrimination is very poor. Finally, in listening to transmission from a broadcasting station, the relatively large changes in modulation due to the rapidily changing intensity of the voice or music tend to mask differences due to changes in transmission.

In Figure 4 is shown a simultaneous record of the radio and audio frequency amplitudes in musical selections transmitted from WBZ, Springfield, Massachusetts, to Newton Centre, Massachusetts, a distance of 118 kilometers (74 miles). The fluctuations in intensity shown in the lower or radio frequency record are practically entirely caused by changes in transmission, whereas the variations shown in the upper or audio frequency record are due to both changes in transmission and in modulation. Thus,

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from 8.33 to 8.35 P. M. the Springfield transmitter was not being modulated, and the audio frequency record remains near its base or zero line. But over this same period the intensity of the electric field is fairly high, as is shown by the lower record. Inasmuch as the galvanometer employed integrates the rectified radio frequency current over periods of several seconds, and as good broadcasting stations are rarely over-modulated, the radio frequency record, if taken in the immediate neighborhood of the station, would be practically a straight horizontal line, as will appear from the next figure.

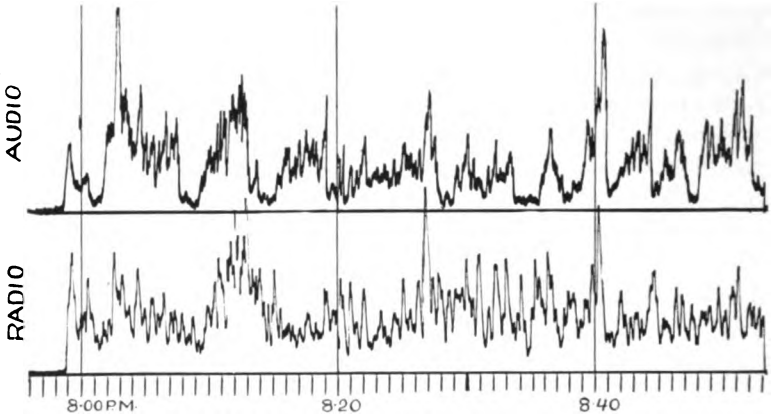

FIGURE 4—WBZ, Springfield, Massachusetts—Newton Centre, Massachusetts, 118 km. (74 miles), April 6, 1923

The first records I shall show are of daytime transmission. In Figure 5 the distance is 11 kilometers (6.9 miles), and the station was then operating at 830 kilocycles (360 meters). The

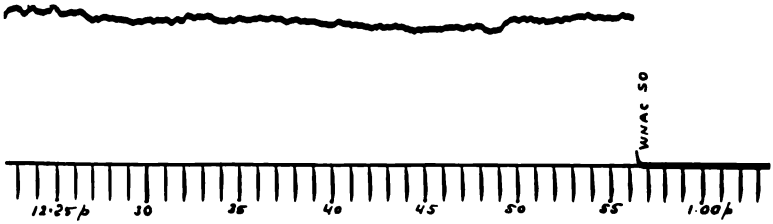

FIGURE 5—WNAC, Boston, Massachusetts—Newton Centre, Massachusetts, 11 km. (6.9 miles), March 1, 1923

more pronounced ripples on this record are undoubtedly due to accidental variations at the transmitter, such as fluctuations in plate voltage due to power line changes, and possibly to occasional

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slight over-modulation. If the broadcasting station radiation meter were of the recording type, it would probably make a very similar record. As to the tiny ripples, which occur several to the minute, I am inclined to attribute some of these to true daytime transmission change, in view of the fact that when the distance is increased, these same ripples appear with increased amplitude, but without material change in character.

In Figure 6 is shown a superposed pair of records made in daytime from the Boston station WNAC at 1,080 kilocycles (278 meters) at two receiving points in Newton Centre, separated 550 meters (1,800 feet). The coincidence is practically perfect, which would indicate that the fluctuations shown are either all at the transmitter, or else that they cover a sufficiently large area to include both receiving points. The method employed for insuring coincidence in time was the very simple and effective one of making a sharp jog in the record line whenever the announcer said "WNAC."

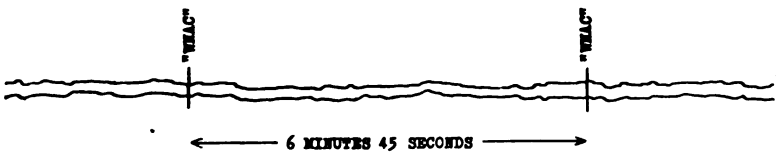

FIGURE 6

This eliminated any possible difference between the watches of the two observers, and enabled the records to be very accurately superposed.

In Figure 7 are shown daytime records from three distant broadcasting stations, at ranges of 60, 290 and 225 kilometers (38, 181, and 141 miles), respectively. These records are typical of average daytime transmission from these distances on moderately disturbed days. The points marked "s s s " on these records are disturbances due to spark working somewhere in the vicinity of Boston, which are, of course, prominent with the rather high amplification necessary for distant day records.

Daytime transmission is, however, different for the same station from day to day, and this is well illustrated in Figure 8, which is of day transmission from WBZ, Springfield, Massachusetts, to Newton Centre, Massachusetts, a distance of 118 kilometers (74 miles). June 9 was clearly a highly disturbed day

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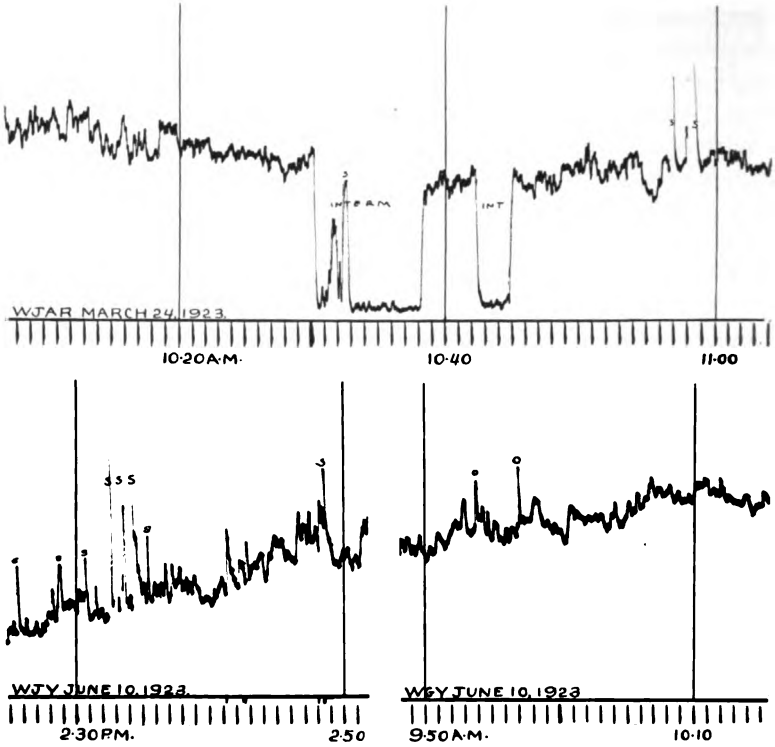

FIGURE 7—Daytime Reception at Newton Center, Massachusetts, from WJAR, Providence, Rhode Island, 60 km. (38 miles), WJY, New York City. 290 km. (181 miles), and from WGY, Schenectady, New York, 225 km. (141 miles)

for radio transmission, June 10 moderate.y disturbed, while June 16 was obviously quiet. These three records fairly represent the average and extreme cases of transmission over this particular range, and altho taken in summer, do not differ from winter daytime transmission.

The normal transition from day to night conditions is well shown in Figure 9, which covers a period of four hours. The late afternoon record beginning in the upper left-hand corner indicates a highly disturbed day (it merely appears flat by comparison with Figure 8 because of its much lower ordinates) with the usual weak field up to about forty minutes before sunset.

Then well-marked large amplitude fluctuations commence, increasing in intensity, and coming in groups with an interval of about 7 minutes. No special disturbance appears to mark exact sunset at either Newton Centre or Springfield, but the amplitude

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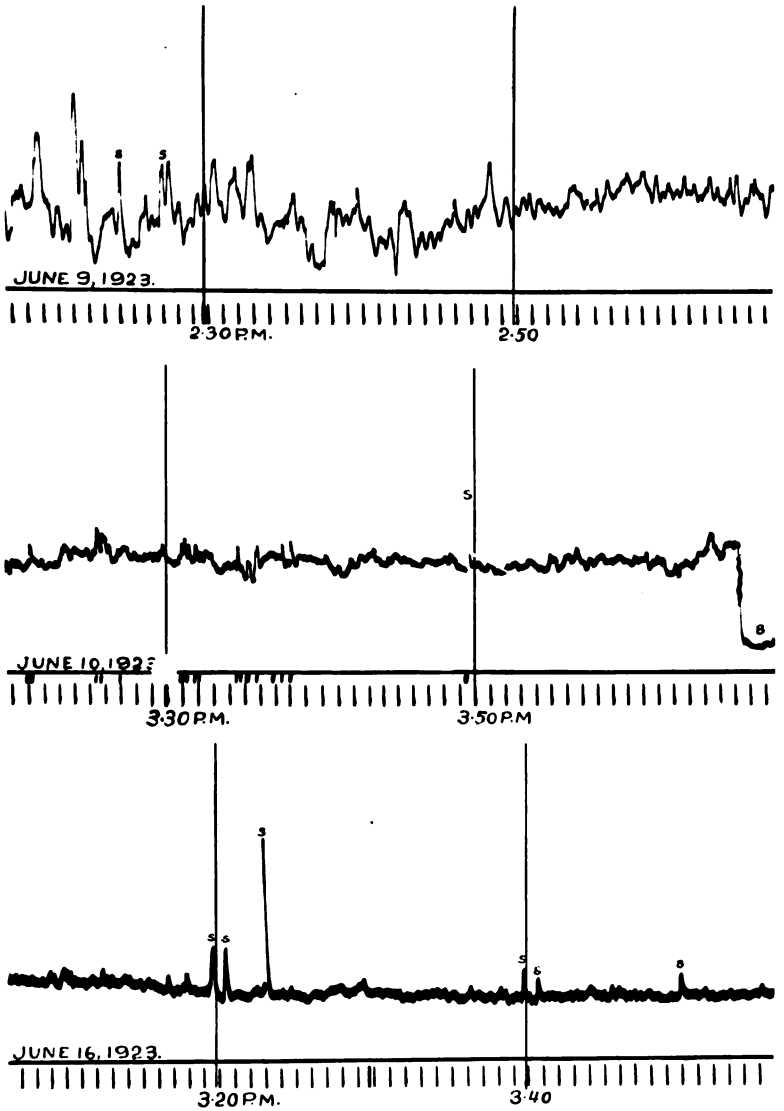

FIGURE 8—WBZ, Springfield, Massachusetts—Newton Centre, Massachusetts, 118 km. (74 miles), Daytime

of the short period fluctuations, and also the mean intensity of the electric field, increases to a maximum about an hour after sunset, and then declines again. This record is typical of many which I have obtained thru the sunset period, and the tendency of the short period oscillations to occur in more or less periodic groups will be shown in many subsequent records.

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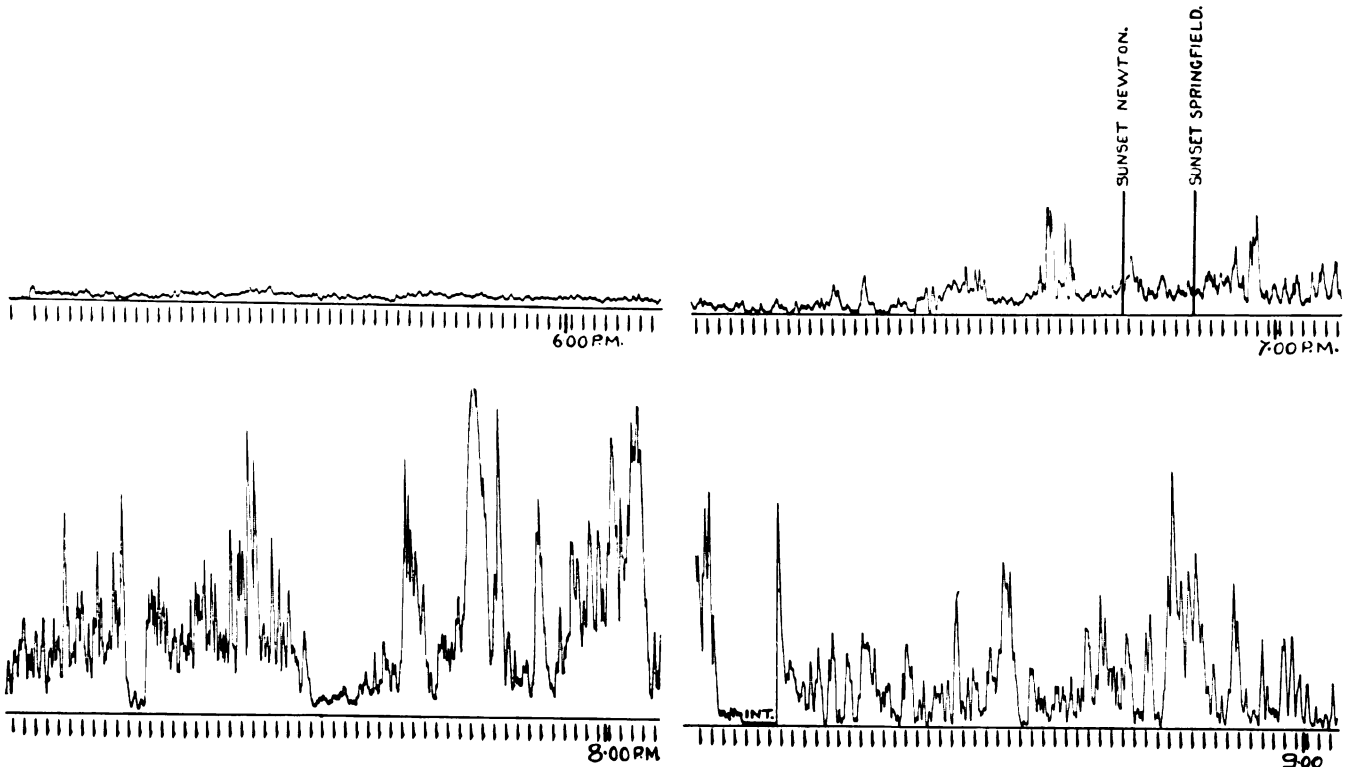

FIGURE 9—WBZ, Springfield, Massachusetts—Newton Centre, Massachusetts, 118 km. (74 miles), Day to Night, August 15, 1923

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Day transmission over short distances, as we have seen from the simultaneous record of Figure 6, does not show any marked differences when recorded at separated points, perhaps because the daytime fluctuations are then too weak to be recorded. But

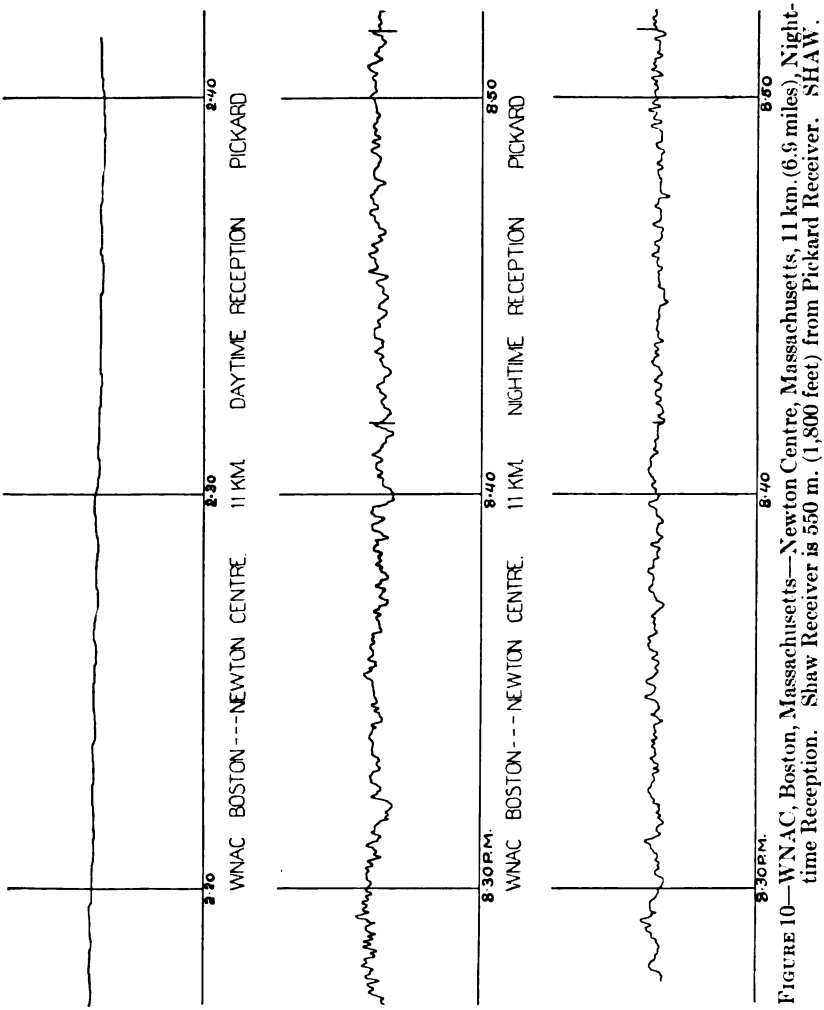

FIGURE 10—WNAC, Boston, Massachusetts—Newton Centre, Massachusetts, 11 km. (6.6 miles), Night-time Reception. Shaw Receiver is 550 m. (1,800 feet) from Pickard Receiver. SHAW.

at night simultaneous records clearly show transmission variations, which are different for even slightly separated receivers. The upper record of Figure 10 is the normal daytime transmission from Boston to Newton Centre, shown merely for comparison, and essentially similar to the records shown in Figure 6. The two

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lower records are of night transmission from the same station in Boston, to two points in Newton Centre, separated by 550 meters (1,800 feet). These records are exactly superposed, the announcer check points appearing at 8.42 and 8.52 P. M. on both records, and it will be immediately apparent that there is little correspondence between them. So far as my analysis of such records has proceeded, the only points of similarity are in the longer swings, which are here very poorly shown because of the flattening caused by the small amplitude and relatively long time scale.

At the left-hand side of Figure 11 are shown five night records of transmission from Boston to Newton Centre, ranging from highly disturbed to quiet. At the right-hand side of the same figure are records made from distant stations on the same nights, and within fifteen minutes of the time of the local station records. It will be seen that on the nights when the record from the local station showed marked fluctuations, the distant stations showed a predominance of short period swings, while on the quiet nights the distant stations gave a predominance of longer period variation.

In Figure 12 are shown four night records of WNAC, Boston, Massachusetts, as received at Scituate, Massachusetts, 33 kilometers (21 miles) distant, of which the first 15 kilometers (9 miles) was over sea water, and the remainder over land. These records also show nights that range from quiet to highly disturbed, and differ in no essential respect from those shown in Figure 11, save for a marked increase in amplitude, caused by the increased distance. This increase in amplitude has, however, made apparent some of the longer periods, which, altho present in Figures 10 and 11, are masked by the flattening.

Figure 13, like the preceding figure, is of night transmission from Boston to Scituate, a range of 33 kilometers (21 miles). It covers a period of nearly three hours, and is a beautiful example of oscillation groups with nearly quiescent periods between. There is a strong resemblance to a seismograph record of a distant and severe earthquake. The change of axis just before 8.30 P. M. is perhaps due to a sudden change of adjustment of the receiving apparatus.

The four records shown in Figure 14 are at a distance of 68 kilometers (42 miles) from station WMAF at South Dartmouth, Massachusetts, to Scituate, Massachusetts, the frequency of WMAF being 845 kilocycles (wave length 355 meters). These four records are fairly representative of the different night con-

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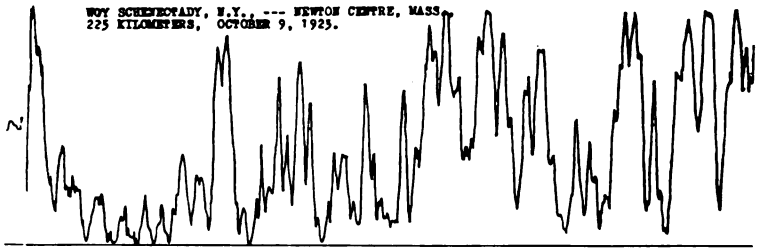

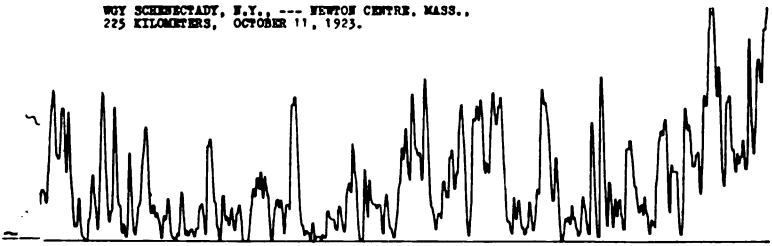

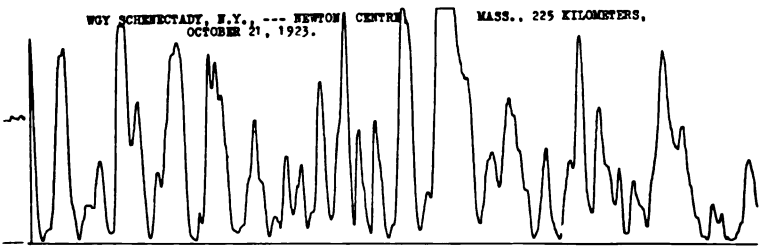

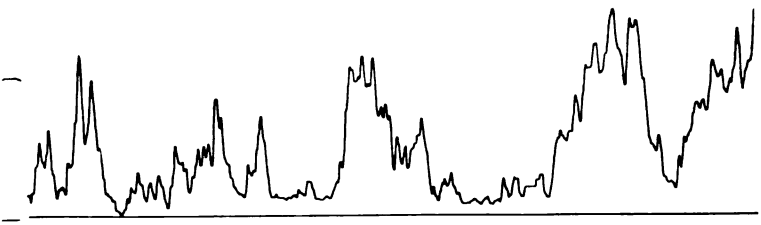

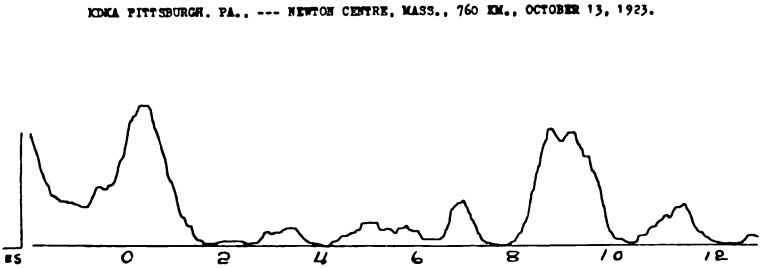

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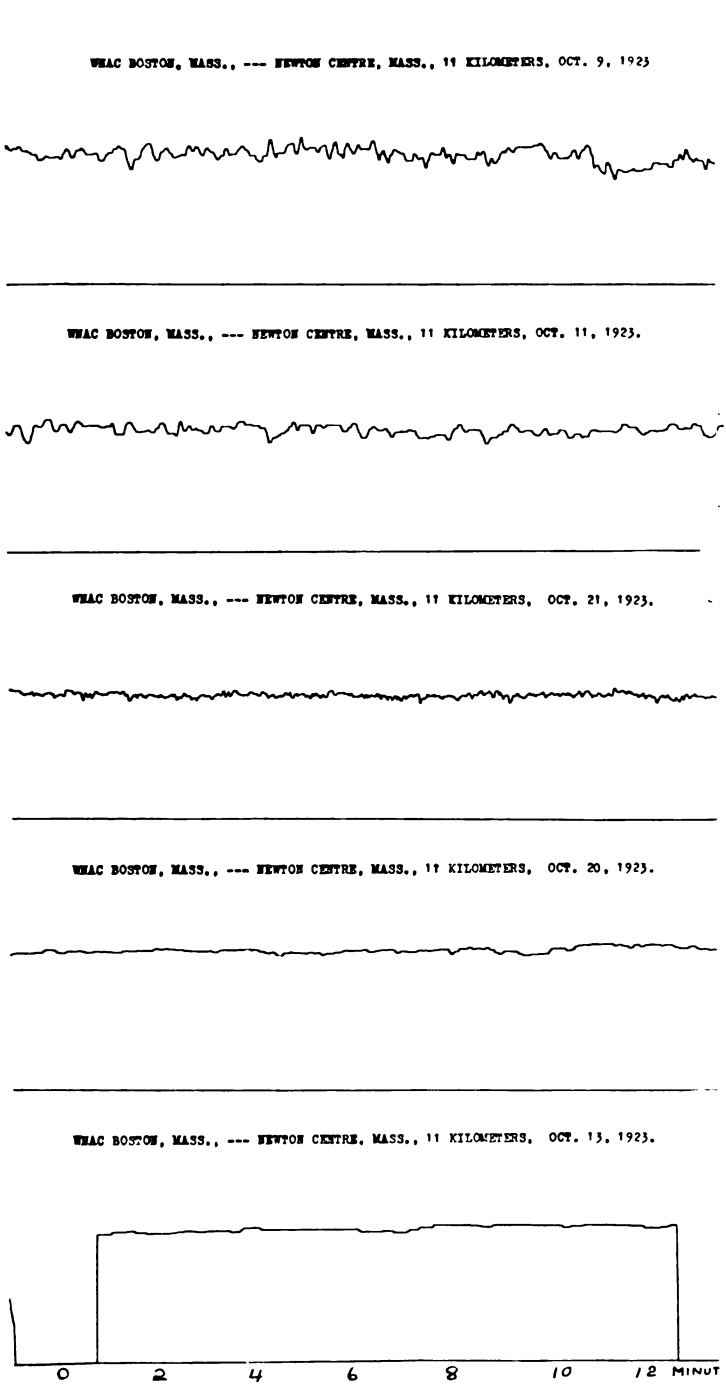
WHAC BOSTON, MASS., --- NEWTON CENTRE, MASS., 11 KILOMETERS, OCT. 9, 1923
WHAC BOSTON, MASS., --- NEWTON CENTRE, MASS., 11 KILOMETERS, OCT. 11, 1923.
WHAC BOSTON, MASS., --- NEWTON CENTRE, MASS., 11 KILOMETERS, OCT. 21, 1923.
WHAC BOSTON, MASS., --- NEWTON CENTRE, MASS., 11 KILOMETERS, OCT. 20, 1923.
WHAC BOSTON, MASS., --- NEWTON CENTRE, MASS., 11 KILOMETERS, OCT. 13, 1923.
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FIGURE 11—Night-time Comparison Reco

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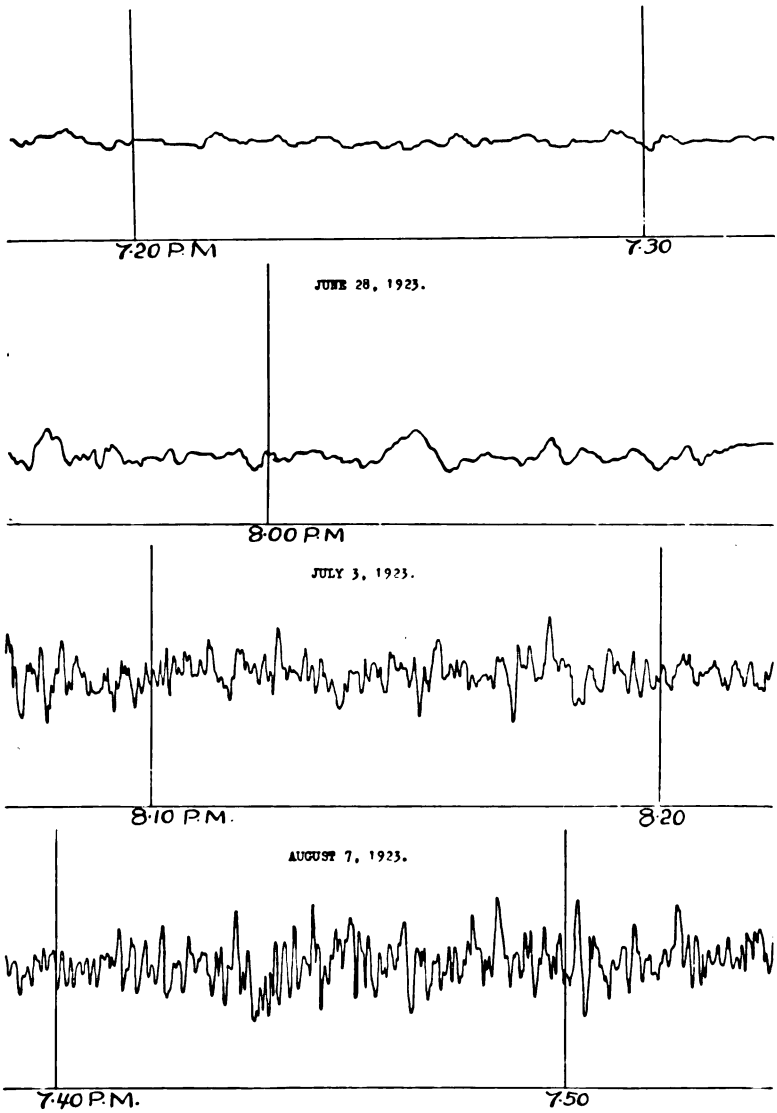

FIGURE 12—WNAC, Boston, Massachusetts—Scituate, Massachusetts, 33 km. (21 miles), of which the first 15 km. (9 miles) are over sea water. July 29, 1923

ditions over this range. They differ from the preceding shorter distance records in that the field no longer fluctuates about a nearly horizontal axis, and also in that the intensity occasionally falls to nearly zero. South Dartmouth, as normally received at Scituate in the daytime, is a strong station with very slight fluctua-

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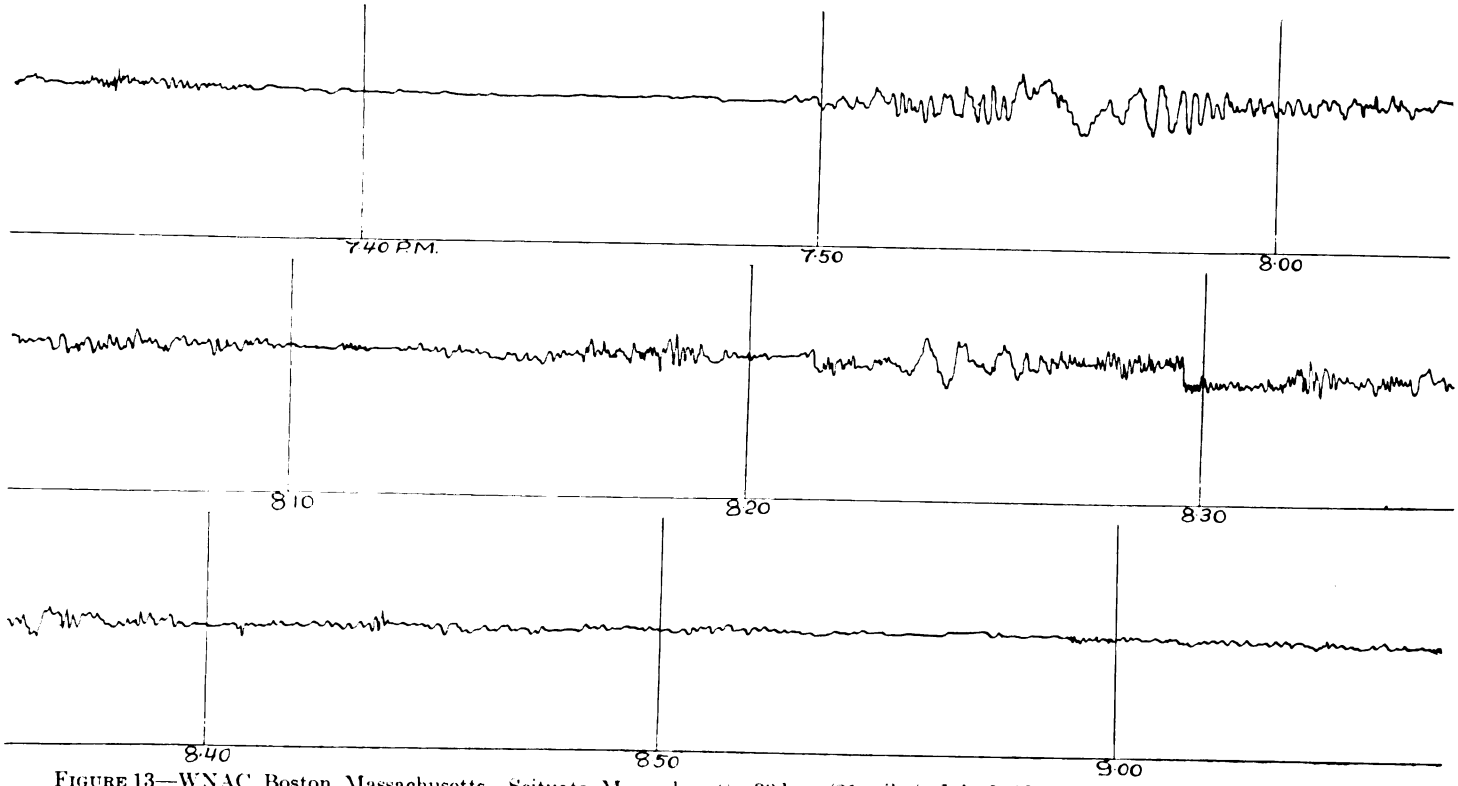

FIGURE 13—WNAC, Boston, Massachusetts—Scituate, Massachusetts, 33 km. (21 miles), July 6, 1923. 15 km. (9 miles) over Sea Water

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tion, whereas at night-time not only does the record show an almost complete extinction at times, but the sound entirely vanishes in the monitoring phones, which shows very conclusively that the night intensity of a distant station may fall far below the daytime intensity.

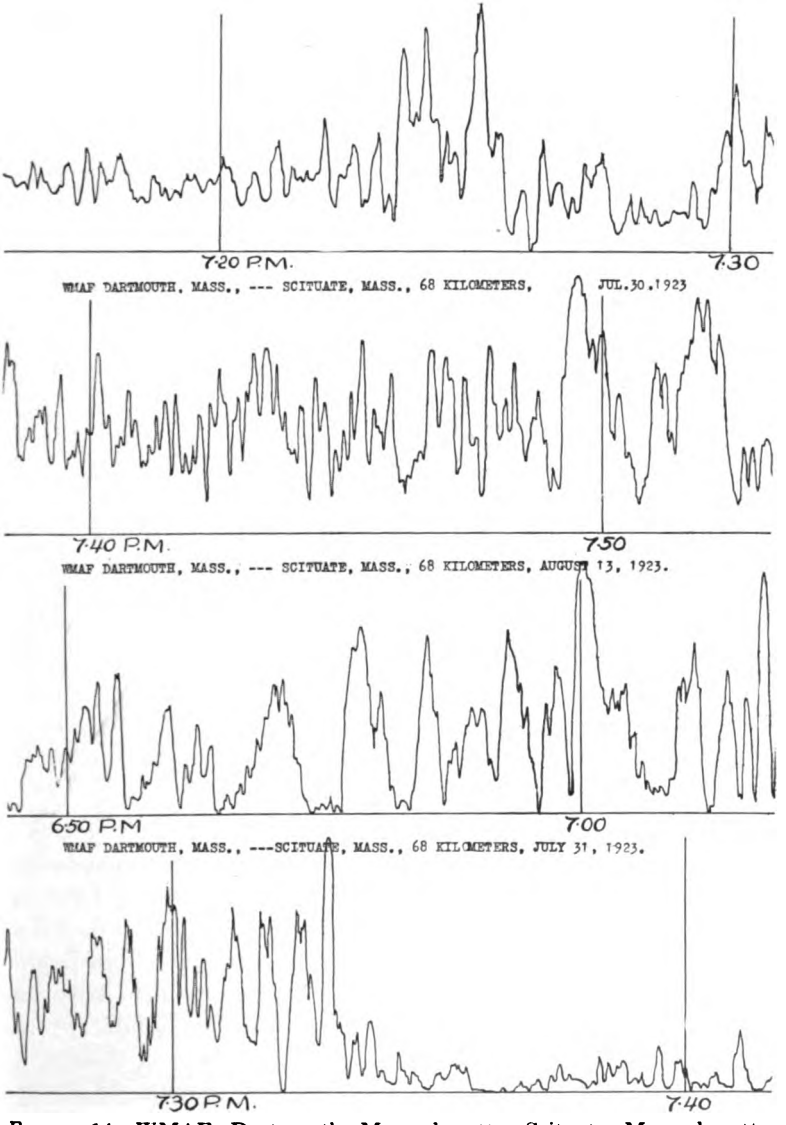

FIGURE 14—WMAF, Dartmouth, Massachusetts—Scituate, Massachusetts, 68 km. (42 miles), July 27, 1923

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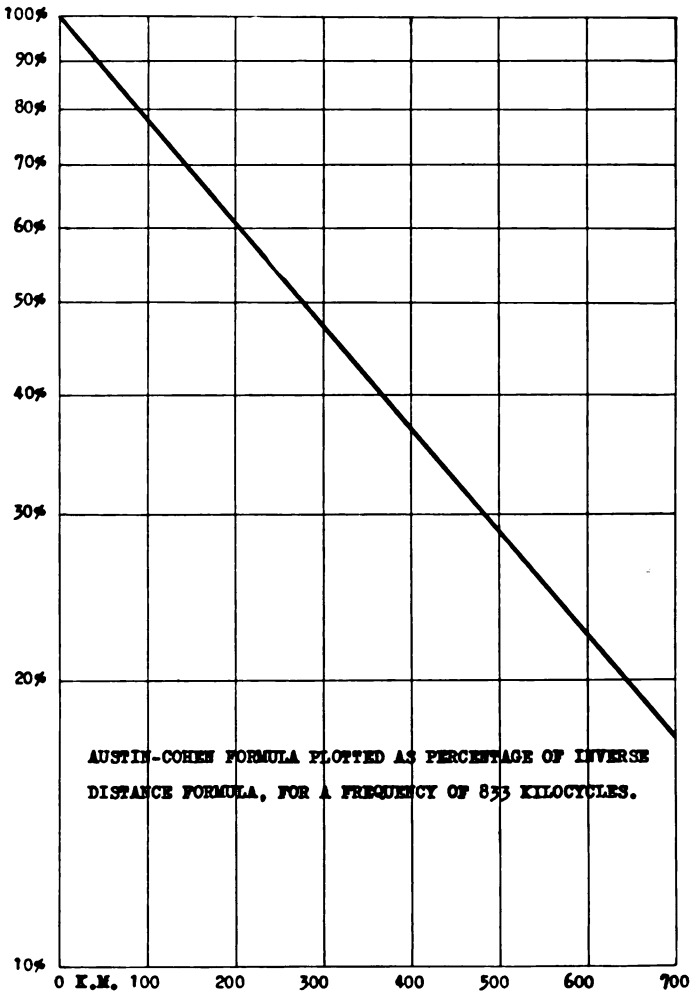

FIGURE 14A

In Figure 14-A I have plotted the Austin-Cohen transmission formula as a percentage of the simple inverse distance formula against distance, for a transmission frequency of 833 kilocycles (360 m.). At 68 kilometers (42 miles), the distance from South Dartmouth to Scituate, the difference in electric field between the two formulas is only about 16 percent, which would appear in the voltage-squared ordinates of my records as a change of some 35 percent. Yet the record shown in Figure 14 shows variations which are at times enormously greater than this. Similarly, as will be seen from succeeding records, the transmis-

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sion from Springfield to Newton Centre, 118 kilometers (78 miles) apart, which according to the difference between the Austin-Cohen and the inverse distance should show only 25 percent variation in electric field, or a change in my ordinates of about 56 percent, actually show fluctuations from minute to minute of tens or hundreds of times. On the other hand, Messrs. Nichols and Espenschied[2A] have found that reception fluctuations in oversea transmission tend to lie between the inverse distance and the Austin-Cohen formula as upper and lower limits. It would appear from this that overland transmission is subject to much greater short period fluctuation than transmission over salt water.

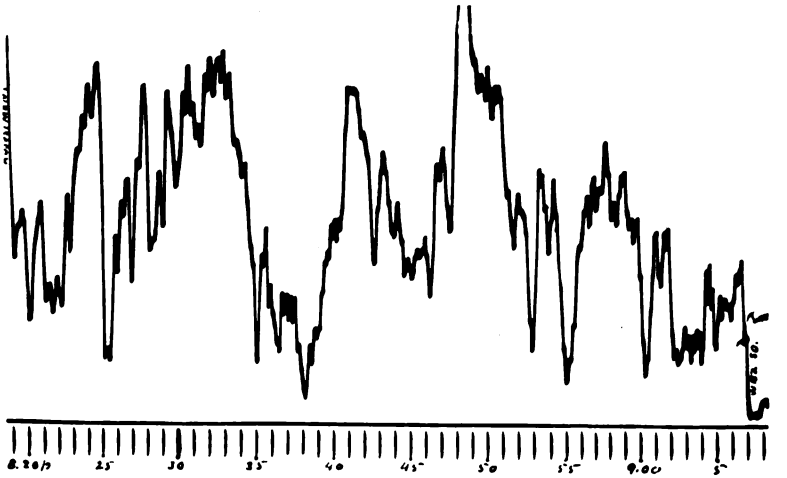

FIGURE 15—WBZ, Springfield, Massachusetts—Newton Centre, Massachusetts, 118 km. (74 miles), February 10, 1923

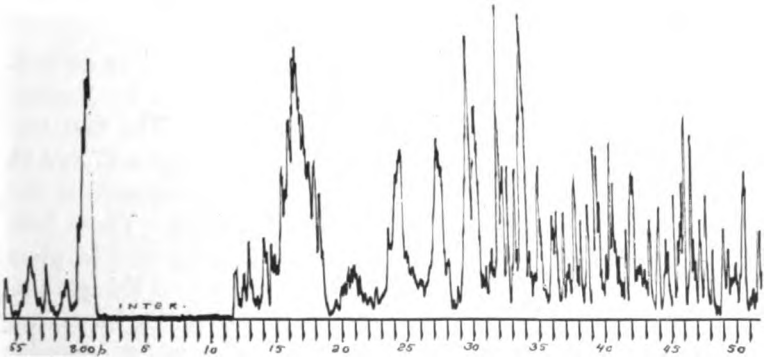

FIGURE 16—WBZ, Springfield, Massachusetts—Newton Centre, Massachusetts, 118 km. (74 miles), April 18, 1923

[2A] "Radio Extension of the Telephone System to Ships at Sea," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, Volume 11, Number 2, June, 1923.

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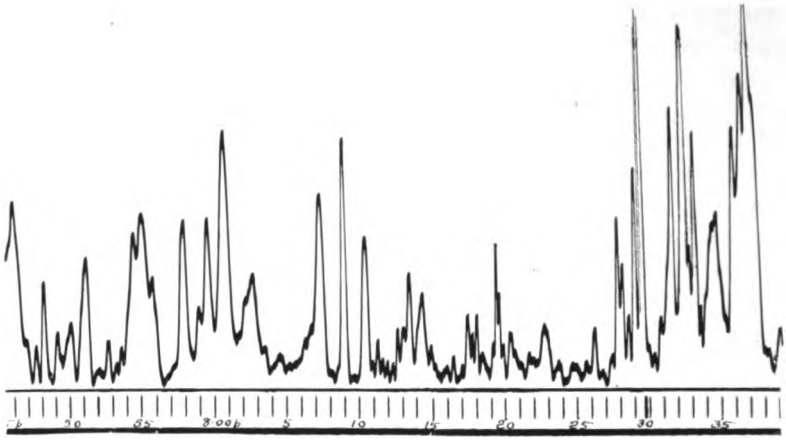

FIGURE 17—WBZ, Springfield, Massachusetts—Newton Centre, Massachusetts, 118 km. (74 miles), June 16, 1923

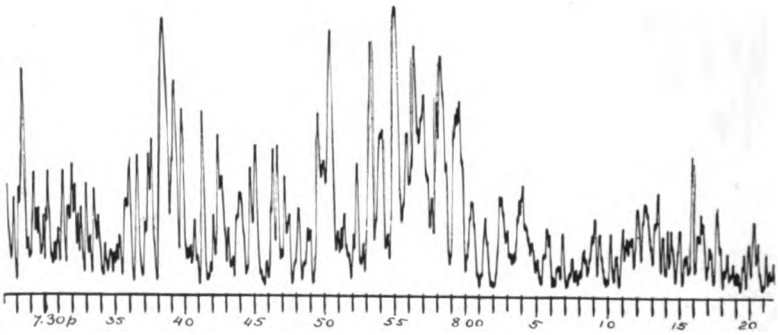

FIGURE 18—WBZ, Springfield, Massachusetts—Newton Centre, Massachusetts, 118 km. (74 miles), September 11, 1923

The four records shown in Figures 15, 16, 17, and 18 are representative of night transmission from Springfield to Newton Centre, a distance of 118 kilometers (74 miles). The first two records are at 750 kilocycles (400 meters), while Figures 17 and 18 are at 890 kilocycles (337 meters), as they are subsequent to the change in frequencies effective on May 15, 1923. These four records run from quiet to highly disturbed nights, and it is interesting in connection with the change in frequency of this station that examination of my numerous records before and after the date of the change does not show any difference which could be attributed to this change.

Figures 19, 20, and 21 are typical of Schenectady to Newton Centre transmission at night, a distance of 225 kilometers (141

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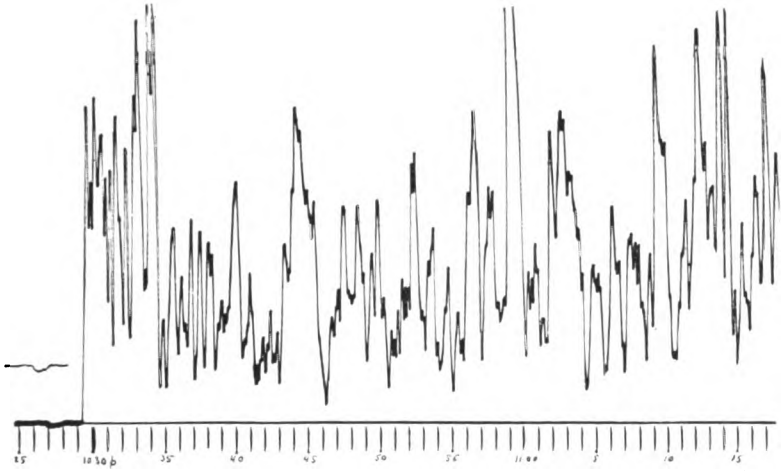

FIGURE 19—WGY, Schenectady, New York—Newton Centre, Massachusetts, 225 km. (141 miles), February 16, 1923

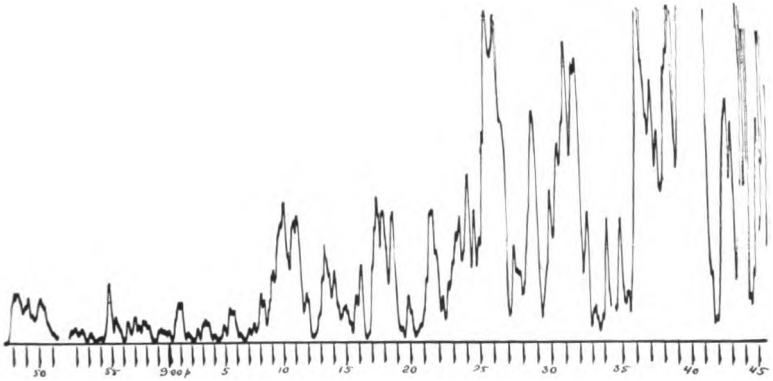

FIGURE 20—WGY, Schenectady, New York—Newton Centre, Massachusetts 225 km. (141 miles), April 26, 1923

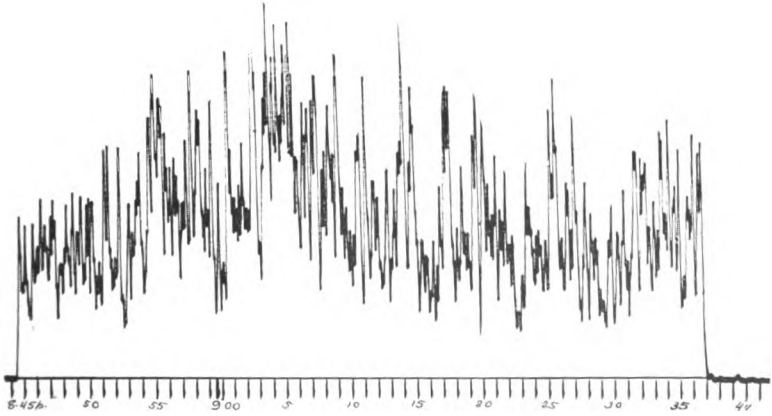

FIGURE 21—WGY, Schenectady, New York—Newton Centre, Massachusetts, 225 km. (141 miles), September 27, 1923

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miles). We have now apparently reached the transmission distance at which the shorter period fluctuations are at their maximum, altho perhaps this may really lie somewhere between Springfield at 118 kilometers (74 miles) and Schenectady at 225 kilometers (141 miles). From now on the records shown will have a predominance of the longer period swings.

Figures 22 and 23 are representative of night transmission from WJZ, New York City, to Newton Centre, a distance of 290 kilometers (181 miles). Figure 22 is of the opening night of Broadcast Central (as duplex stations WJY and WJZ are termed), and the interest taken in this by my Newton neighbors is shown by the "squeal" record appearing just above the time scale markings at the lower edge of the sheet. Each time a beat note from a neighboring oscillating receiver was heard in the monitoring telephones, a spot of light was flashed on the record sheet. These squeals are noticeably grouped most closely around low spots in the transmission, and also around any interval when the station was not modulating. It is a fixed idea of the average broadcast listener that when signals are weak or absent his receiver must be suspected and adjusted. At this distance of transmission the character of the record begins to change, and the longer period swings, measured in minutes, begin to become more prominent.

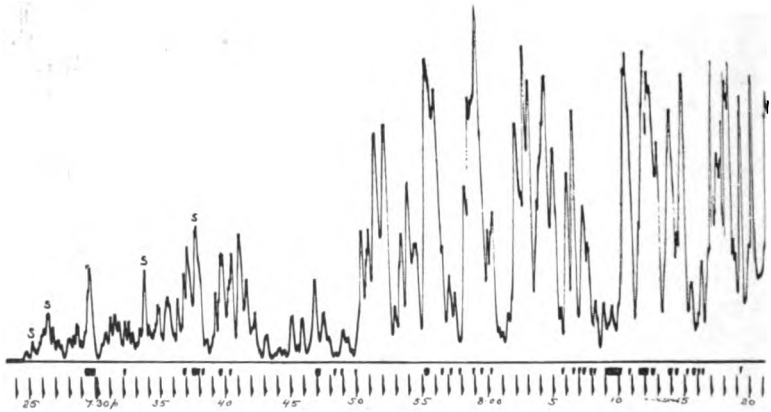

FIGURE 22—WJZ, New York City—Newton Centre, Massachusetts, 290 km. (181 miles), May 15, 1923

Figures 24 and 25 are KDKA, Pittsburgh, Pennsylvania, as received by night at Newton Centre, a distance of 760 kilometers (475 miles). The distance is now great enough to bring out strongly the long period swings, of about five and a half minutes in Figure 24, and about ten minutes in Figure 25.

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On quiet evenings, as will be seen from later records of the same station, the long periods are even more in evidence.

This record is typical of night transmission from WCAP, Washington, District of Columbia, to Newton Centre, a distance of 630 kilometers (394 miles). Altho the distance is less than that

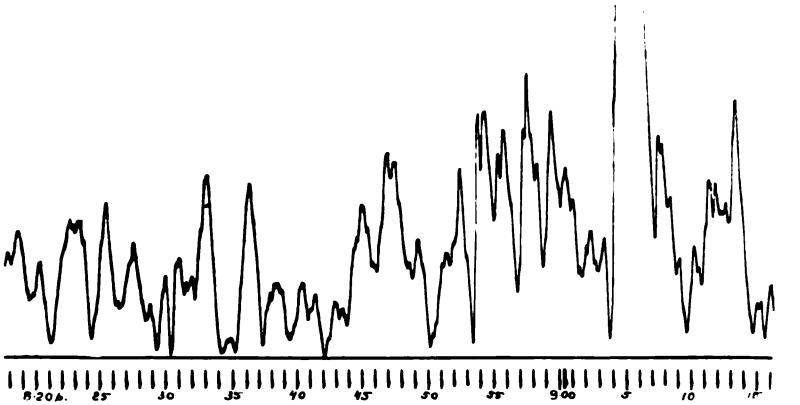

FIGURE 23—WJZ, New York City—Newton Centre, Massachusetts, 290 km. (181 miles), September 16, 1923

of the records shown in Figures 24 and 25, the transmission from this station is always relatively free of fluctuations having a period of one minute or less. It would seem that the ground route followed by the transmission played some part in determining the character of the variations.

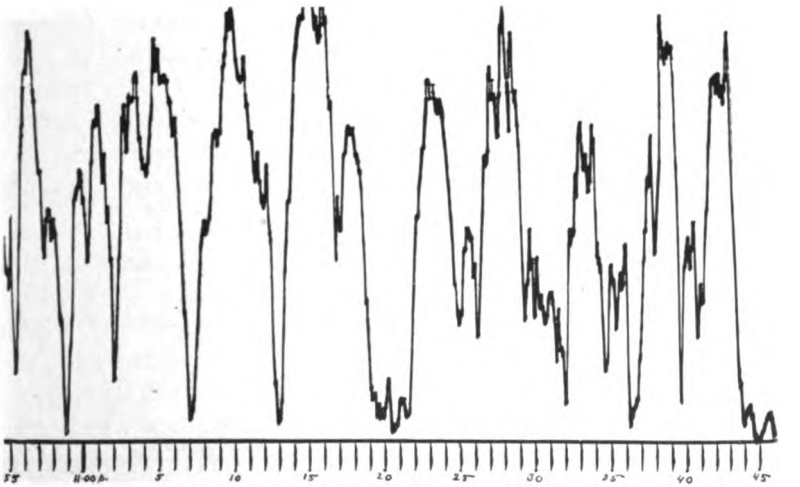

FIGURE 24—KDKA, Pittsburgh, Pennsylvania—Newton Centre, Massachusetts, 760 km. (475 miles), February, 14, 1923

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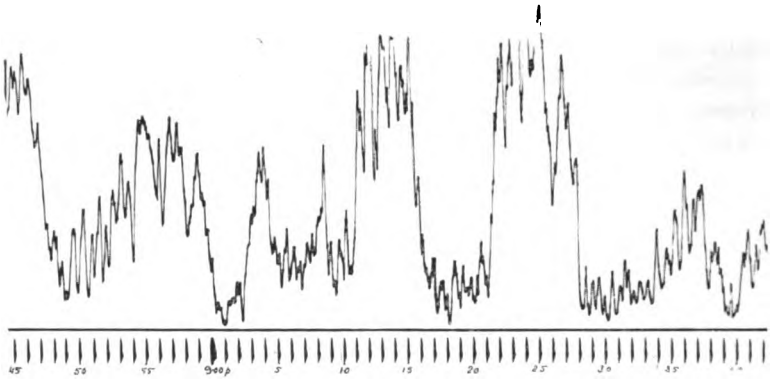

FIGURE 25—KDKA, Pittsburgh, Pennsylvania—Newton Centre, Massachusetts, 760 km. (475 miles), March 22, 1923

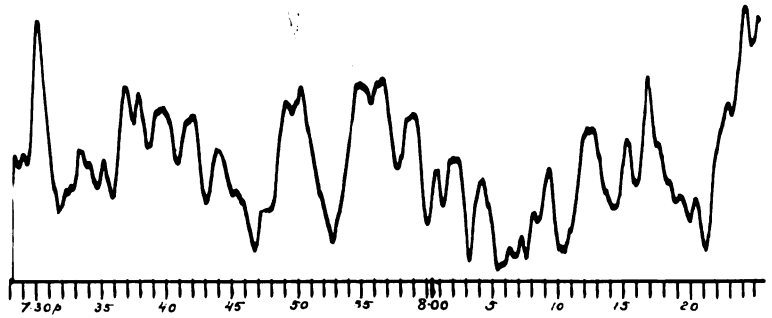

FIGURE 26—WCAP, Washington, District of Columbia—Newton Centre, Massachusetts, 630 km. (394 miles), September 1, 1923

Figure 27 is of night transmission from WDAP, Chicago Illinois, to Newton Centre, Massachusetts, a distance of 1,400 kilometers (875 miles). A long swing of about twenty minutes is clearly brought out in this record, with superimposed periods of about thirty and sixty seconds. This record was made on a moderately disturbed night; on a quiet night the curve is much smoother.

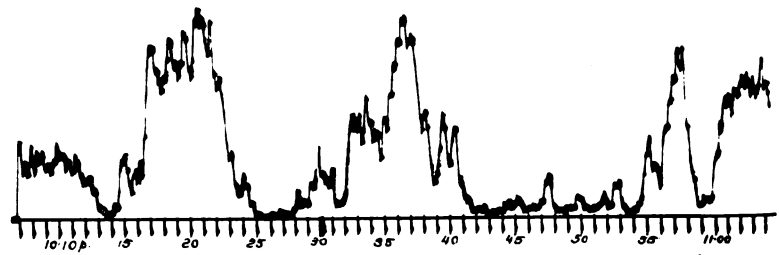

FIGURE 27—WDAP, Chicago, Illinois—Newton Centre, Massachusetts, 1,400 km. (875 miles), August 2, 1923

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Figure 28 is of night transmission from WBAP, Fort Worth, Texas, to Newton Centre, Massachusetts, a distance of 2,600 kilometers (1,630 miles). A long period of some fifteen minutes is prominent, and with the rather high amplification necessary for this distant station, the background disturbance becomes quite appreciable.

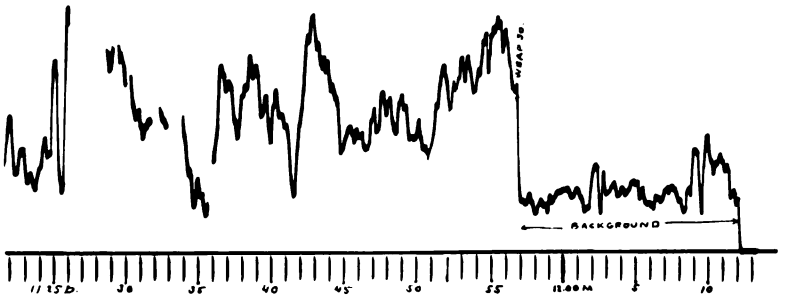

FIGURE 28—WBAP, Fort Worth, Texas—Newton Centre, Massachusetts, 2,600 km. (1,620 miles), February 8, 1923

Figure 29 is a simultaneous night record at Newton Centre of Schenectady, distant 225 kilometers (141 miles), taken at receiving points separated 550 meters (1,800 feet). With the exception of the period of high intensity at about 8.10 P. M., it is difficult to identify any of the detail of one record with that of the other. The longer period swings of these records are, however, substantially the same, and this becomes apparent if the curves are smoothed.

The fact that altho the shorter period swings are not the same on records made at slightly separated points, the longer period fluctuations are substantially identical, is well brought out by Figure 30, which is of night transmission from Pittsburgh, Pennsylvania to Newton. This was made on a quiet evening, so that the short period variations are weak, and altho considerable distortion occurs, particularly around 8.50 P. M., the main features of the two records are the same.

If, however, the receiving points are at all widely separated, there is little or no similarity between their records of a distant station. In Figure 31 the transmitting station is WBZ, Springfield, Massachusetts, and the two receiving points are Newton Centre and Scituate, Massachusetts, 118 and 155 kilometers (74 and 97 miles), respectively, from Springfield, and 41 kilometers (26 miles) from each other. Aside from the general increase in intensity after sunset, which was at 7.20 P. M., there is no correspondence between these two records.

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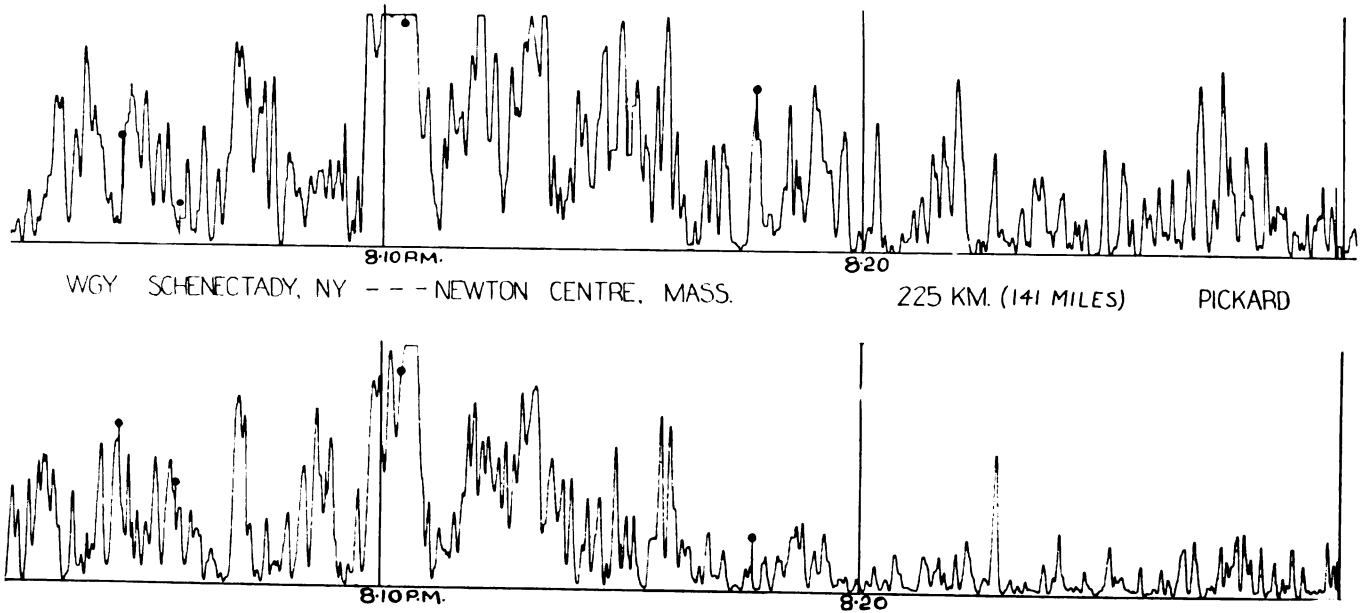

FIGURE 29—WGY, Schenectady, New York—Newton Centre, Massachusetts, 225 km. (141 miles)

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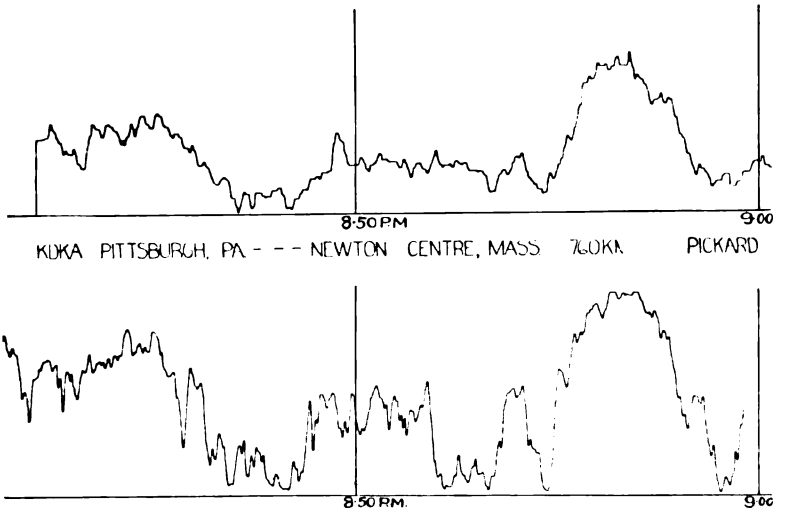

FIGURE 30—KDKA, Pittsburgh, Pennsylvania—Newton Centre, Massachusetts (760 km. (475 miles), SHAW

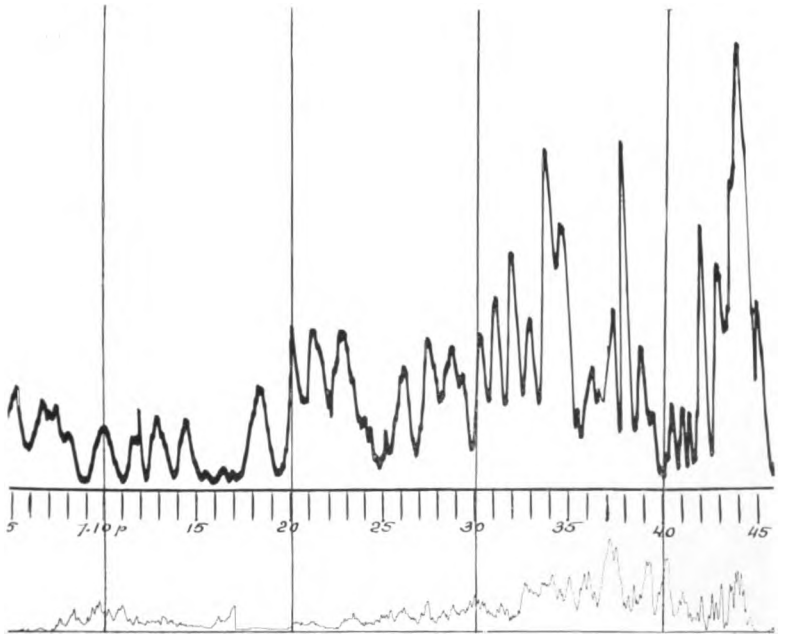

FIGURE 31—Simultaneous Reception of WBZ, Springfield, Massachusetts, at Newton Centre, Massachusetts, and Scituate, Massachusetts, Distant 118 and 155 km. (74 and 97 miles). Scituate is 41 km. (26 miles) from Newton Centre. June 9, 1923

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Figure 32 is a simultaneous night record at the same receiving point in Newton Centre, of two stations in Newark, New Jersey, distant 315 kilometers (197 miles), and, of course, operating at different frequencies. There are no points of correspondence between these records, save accidental coincidence of maxima and minima which would occur in the superposition of any two random complex curves.

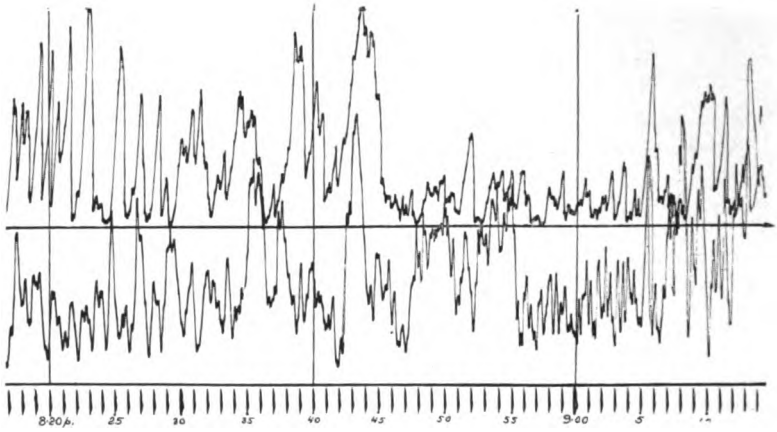

FIGURE 32—Simultaneous Record at Newton Centre, Massachusetts, of WOR and WJZ (original location), both at Newark New Jersey, Distant 315 km. (197 miles), March 2, 1923

Figures 33, 34, and 35 illustrate simultaneous transmission to Newton Centre from widely separated broadcasting stations. As in Figure 32, there is a complete lack of correspondence.

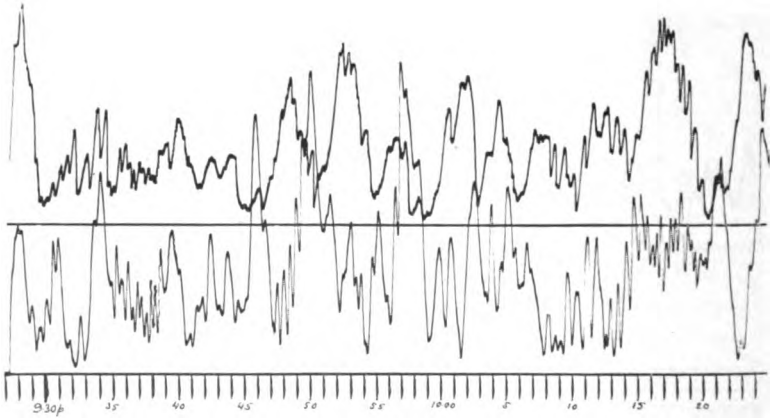

FIGURE 33—Simultaneous Record at Newton Centre, Massachusetts, of WEAF, New York City, and WJZ, Newark, New Jersey, Distant 290 and 315 km. (181 and 197 miles), March 10, 1923

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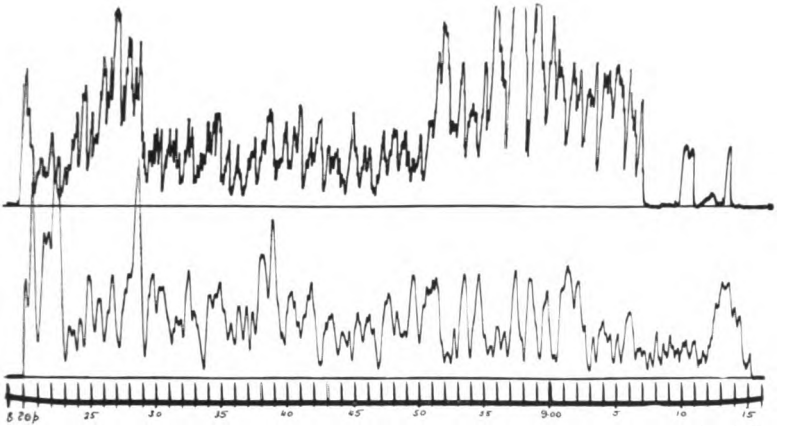

FIGURE 34—Simultaneous Record at Newton Centre, Massachusetts, of WBZ, Springfield, Massachusetts, and WEAF, New York City, Distant 118 and 290 km. (74 and 181 miles), February 24, 1923

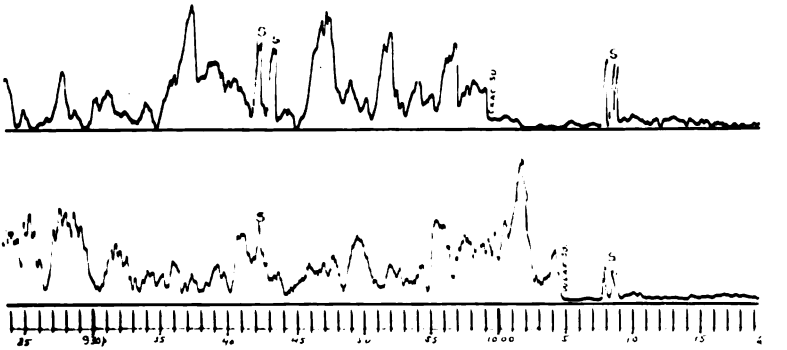

FIGURE 35—Simultaneous Record at Newton Centre, Massachusetts, of CKAC, Montreal, Canada, and WEAF, New York City, Distant 405 and 290 km. (254 and 181 miles), February 24, 1923

In Figure 36, which is a simultaneous night record at Newton Centre from New York City and Springfield, there is one correspondence; namely, the dead period beginning about 8.42 P. M. Unfortunately for further possible points of similarity, Springfield signed off at 9.04 P. M.

I have said above that reception from a distant station is not the same at separated points, unless the separation is very small indeed. In addition to extremely small separation, it is also necessary for identical records that the two antennas be of the same directional properties, and similarly oriented.

In Figure 36-A is shown a simultaneous record on two open antennas, separated only 30 meters (97 feet) on centers. The

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two records are substantially identical, and the slight differences between them are probably due to the fact that one of the open antennas was a large T, while the other was of the inverted L type, so that there was a slight difference in directional properties.

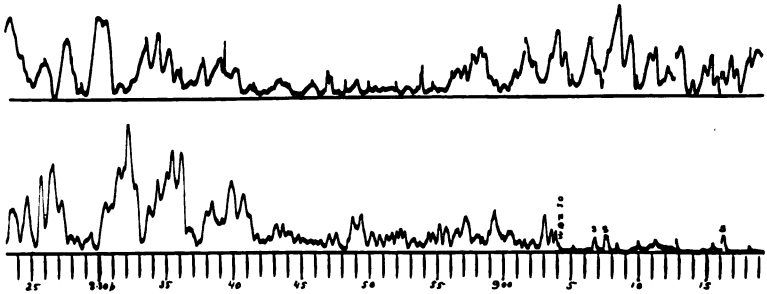

FIGURE 36—Simultaneous Record at Newton Centre, Massachusetts, of WEAF, New York City, and WBZ, Springfield, Massachusetts, Distant 290 and 118 km. (181 and 74 miles), March 10, 1923

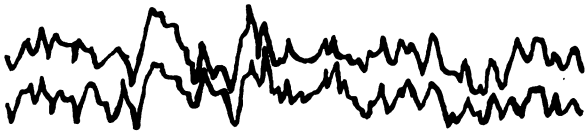

FIGURE 36A—WNAC, Boston, Massachusetts, received at Newton Centre, Massachusetts, 11 kilometers (7 miles) distant, November 13, 1923, on antennas 30 meters (97 feet) apart on centers. Upper record on large open antenna by Shaw; lower record on small open antenna by Pickard. Record is 4 minutes and 40 seconds long

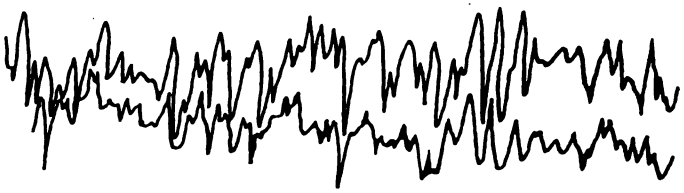

FIGURE 36B—WNAC, Boston, Massachusetts, received at Newton Centre, Massachusetts, 11 kilometers (7 miles) distant, on November 13, 1923, on two small loops, 3 meters (9.7 feet) apart on centers and in line with WNAC. Upper record by Pickard; lower record by Shaw. Record is 6 minutes, 20 seconds

Figure 36-B shows a simultaneous record on two small loops, 3 meters (9.7 feet) apart on centers, and each loop in line with the broadcasting station. Here, also, the records are substantially identical, the slight differences which appear being probably due to differences in the open antenna effect of the two loops, which were compensated.

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But in Figure 36-C, which is a simultaneous record made on antennas of different directional properties, the upper record that of a T open antenna, and the lower on a small loop, separated 15 meters (48 feet) on centers, a total dissimilarity appears, both in shape and amplitude, the loop record showing about four times the amplitude of the open record.

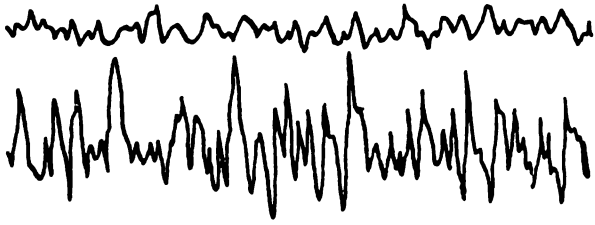

FIGURE 36C—WNAC, Boston, Massachusetts, received at Newton Centre, Massachusetts, 11 kikometers (7 miles) distant, November 13, 1923, on antennas 15 meters (48 feet) apart. Upper record on open antenna by Pickard; lower record on loop by Shaw. Record is 5 minutes long

Figure 36-D is a double record made on two small loops, uncompensated, 3 meters (9.7 feet) apart on centers, at right angles with each other and each loop at 45° from the plane, including the broadcasting station. Here, also, the two records are entirely dissimilar.

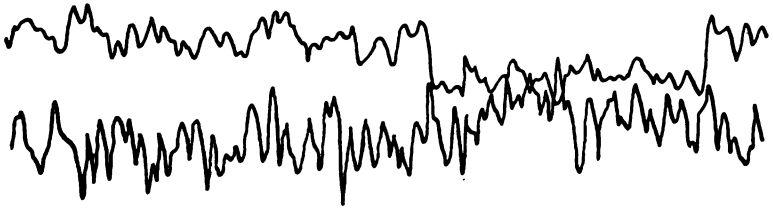

FIGURE 36D—WNAC, Boston, Massachusetts, received at Newton Centre, Massachusetts, 11 kilometers (7 miles) distant, November 13, 1923, on two small loops, 3 meters (9.7 feet) apart on centers. These two loops were at right angles with each other, and at 45° with the radio bearipg of the Boston Station. The upper record was made by Pickard on a loop in a NE-SW plane; the lower record is by Shaw on another loop in a NW-SE plane. Record is 9 minutes long

It has been thought that reception variations in loop reception were largely due to changes in the direction of wave-front, and this may be true in some localities, and with certain distant stations. For example, I have often found large short-period variations in the apparent bearing of ship stations at night, when the receiving point was near the water's edge, and the line of

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transmission ran grazingly along the shore. Under such conditions an apparently weakened or lost signal could sometimes be restored by swinging the loop thru 90°. But inland, and particularly at my principal receiving point in Newton Centre, overland reception from broadcasting stations does not show large changes in wave-front. A loop null point taken on WNAC, Boston, Massachusetts, 11 kilometers (7 miles) distant, remains constant thru an entire evening to within a degree or two, so that the effects shown in the preceding figures can hardly be due to changes in wave-front.

Altho the relation of these short period flucuations to transmission frequency is a matter which I must reserve for a later paper, I have shown in Figure 36E an interesting simultaneous record, made at Newton Centre on two open antennas separated 30 meters (97 feet) on centers, of the two transmission frequencies KDKA, Pittsburgh, Pennsylvania. The upper record is at 920

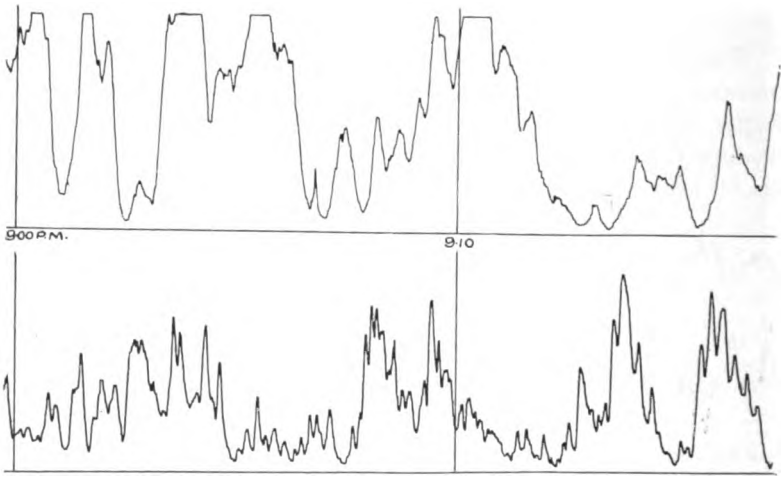

FIGURE 36E—KDKA, Pittsburgh, Pennsylvania, received at Newton Centre, Massachusetts, 760 kilometers (565 miles) November 28, 1923. Upper record 920 kilocycles (326 miles); lower record 3,200 kilocycles (94 miles)

kilocycles (326 meters), while the lower one is at 3,200 kilocycles (94 meters). I selected this particular record because it was out of the ordinary in that there is a general resemblance between the curves; ususlly reception at 3,200 kilocycles is subject to a much more rapid variation than is here shown. Figure 36F, which is a four-hour record of KDKA at 3,200 kilocycles, is more nearly representative of this transmission.

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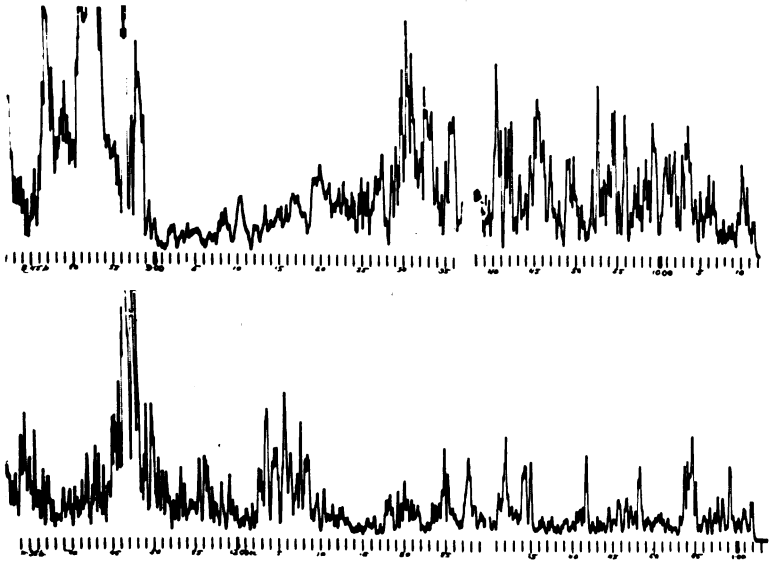

FIGURE 36F—KDKA, Pittsburgh, Pennsylvania, received at Newton Centre, Massachusetts, 760 kilometers (565 miles), 3,200 kilocycles (94 meters), November 29-30, 1923

I have devoted some little study to KDKA's 3,200-kilocycle transmission, and the result, at least in so far as Newton Centre reception is concerned, may be briefly summed up as follows: In the afternoon and early evening reception from KDKA is often better at 3,200 kilocycles than at 920, while later in the evening the lower frequency is usually better. Until recently the higher frequency was quite free of disturbances; now there is a noticeable amount of amateur working at or near this frequency, and a sufficient number of oscillating receivers to much the quality badly at times.

It has probably been observed by all broadcast listeners that in general the quality of reception from a distant station is markedly inferior to that obtained from a local station, this inferior quality being a matter entirely aside from impairmant of reception due to disturbances. I am now convinced that the principal cause of the distortion in reception from distant stations is that the short period variations in transmission are not the same for the carrier wave and its two side bands, and I hope to present in another paper certain records which show this effect. Fortunately there are two obvious remedies for this form of distortion; the use of a single side band for transmission, and the employment of much higher transmission frequencies. Neither

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of these remedies will be immediately popular with the average broadcast listener.

So far as the varying intensity of the sound in the telephone receiver is concerned, there are several simple expedients which will markedly smooth out reception from a distant station. Thus, the grids in a radio frequency amplifier train may be floated on a fairly large condenser, shunted by a high resistance. When reception is weak, the grids assume a small negative potential, and amplification is at a maximum. When the input rises, a a large negative charge is built up of the grids, and amplification is reduced. Similarly, a separate rectifier connected to the output end of the radio frequency amplifier and to the grids will have the same effect.

So far I have confined myself strictly to my records, which have shown you graphically the facts of short period variations in broadcast transmission. There is as yet no adequate theory as to their cause. But I am sure you will agree with me as to the utility of hypotheses. Scientific progress is literally milestoned with hypotheses, most of which are to-day gravestones. But an hypothesis is at least something concrete, and if it stands the attack of new facts, and the further analysis and correlation of our present knowledge, it may form a stepping-stone to the truth. I have already mentioned a further paper on this subject, and in this I hope to present, among other things, the curious results of harmonic analysis of these record curves, some attempted correlation between these fluctuations and other terrestrial happenings, and simultaneous records of field intensity at a large number of separated points. I also hope to establish rather definitely the size and shape of the area within which these variations are identical. But at the present time I feel very strongly that any attempt at a detailed hypothesis would be decidedly premature, altho some rather general conclusions may not be out of place.

A very odd explanation of these variations has been gaining much vogue in the popular scientific press, to the effect that they are due to other receivers in the neighborhood, particularly if these happen to be of the much maligned single circuit regenerative variety. I have many records of transmission, principally made by audibility meter measurement, which run back over fifteen years, when the density of receivers was far less than at present. These records show exactly the same short period fluctuations that now exist, so that this explanation does not seem very plausible. However, I have recently made a number of

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records of reception from distant stations under conditions of severe exposure to nearby regenerative receivers, of which Figure 37 is typical.

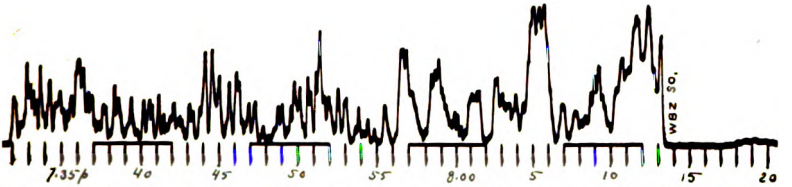

FIGURE 37—WBZ, Springfield, Massachusetts, Received at Newton Centre, Massachusetts, Distant 118 km. (74 miles). Base lines as at 7.37 to 7.42 P.M. Indicate Periods of 5 Minutes each when Single Circuit Regenerative Receiver in Same House and with Adjacent Antenna was in Operation. September 10, 1923

In this record, the broken heavy base line indicates periods when a single circuit regenerative receiver in the same house with my recording set, and with a fair-sized open antenna adjacent to the one employed with the recording set, was in operation. Full regeneration was used almost to the point of oscillation, but this record, like several others which I have made with different exposures of antennas, shows no effect whatsoever.

The fluctuations in radio transmission, at least those of short period, are at first thought most readily explained by the assumption of plural transmission paths; that is, the effect appears to be one of interference. I believe that De Forest was the first to suggest this in 1913,[3] for short period variations in nighttime transmission from the arc stations of the Federal Telegraph Company on the Pacific coast. Such plural paths may lie in altitude or azimuth, or both, but the favorite explanation has been that reflection takes place at some high level in the atmosphere, because of an ionized stratum called "the Heaviside layer."

In order to deflect electromagnetic waves, we may reflect them from a conducting surface or mirror, refract them by a change in optical density of the medium, or bend them by a two- of three-dimensional periodic structure of absorbing and non-absorbing elements. Inasmuch as, in our atmosphere, all three of these methods call for ionization, and as the maintenance of an ionized state requires energy, it is easy to arrange these methods in decreasing order of energy required to produce the observed effects, which would also seem to be the order of increasing probability.

[3] PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 1, number 1, 1913, pages 42-51.

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Maxwell's theory for the relation between the reflecting power of a mirror and its electrical conductivity[4] requires for the effective reflection of the waves employed in broadcasting a conductivity somewhere between fresh water and sea water. Eccles has computed[5] the conductivity necessary for refraction of such waves to be out of the order of one-millionth that of sea water. I have approximately calculated the ionization required to reflect the waves on the assumption of a periodic structure; a sort of "mackerel sky"* of ionized clouds, with the result that a considerably smaller number of ions is required than for refraction by a uniform state of ionization. Because of the smaller amount of energy required to maintain such a stratified state, I have considered this the more probable upper-level condition capable of causing reflection-like effects.

Inasmuch as I have found well-defined short period variations of more than ten percent amplitude in transmission over a distance as short as eleven kilometers (6.9 miles), in which the direct path from transmitter to receiver was at least ten times shorter than the total "reflected" path, it would seem to involve nearly normal reflection with an efficiency of one hundred percent or over, on the conventional Heaviside layer hypothesis. If some recent[6] revisions of our ideas as to the density of our atmosphere at high levels are correct, the Heaviside layer is perhaps twice as high as we have supposed, which apparently removes it as a possible explanation of fluctuations over such short distances of transmission.

In view of the fact that night-time transmission over greater distances than eleven kilometers (6.9 miles) not infrequently falls to much lower levels than at any time during the day, it seems necessary occasionally to invoke the assistance of some form of plural path transmission with its consequent interference effects. But the general tendency of night-time field values is to range between the normal daytime transmission value as a lower limit, and the value given by the transmission formula without the exponential or so-called absorption term as an upper limit. It would seem, therefore, that the principal factor affecting radio transmission was absorption, with reflection-like effects present,but playing a minor role.

[4] Drude, "Physik des Aethers," page 574.

[5] Eccles, "London Electrician," 69, pages 1015-1016, 1912.

*A "mackerel sky" is a dappled cloud formation consisting of heaped-up or rounded clouds arranged in wisps or streamers—EDITOR.

[6] Dobson, "The Characteristics of the Atmosphere up to 200 Kilometers as Indicated by Observations of Meteors," "Quarterly Journal of the Royal Meteorological Society," volume XLIX, 207, July, 1923.

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It may at first seem altogether unlikely that such rapid fluctuations as I have found in transmission could be caused by changes in absorption. Absorption can only be explained as due to ionization, and while it does not seem impossible that ionization changes large enough to affect our whole atmosphere might have a diurnal period and so cause the well-marked difference between day and night transmission, it is difficult to imagine such large scale effects varying with periods of minutes and even seconds. One difficulty in the past has perhaps been our tendency to attribute too much of the ionization to the effect of direct sunlight, and too little to the effect of alpha particles shot out by the sun, which continually fall in drifting clouds into our atmosphere. These charged particles may arrive intermittently and at short intervals, perhaps in groups of small, cloud-like masses, and before the outer limit of our atmosphere is reached the earth's magnetic field begins to deflect and comb them out along its own direction. This process continues, perhaps, down to an elevation of some 150 kilometers (94 miles) or less, eventually drawing these clouds out into long streamers in a south to north direction. Then, perhaps, these streamers are captured by the high level air currents, which over the United States at heights of between 100 and 20 kilometers (63 and 13 miles) drift slowly from west to east, and, finally, they become subject to the stronger and varying low level winds.

I have attempted to illustrate this idea in Figure 38. From *A* to *B* the clouds of charged particles are drawn out in a south-to-north direction as they fall; from *B* to *C* they are deflected at nearly right angles to this course, or in a west-to-east direction, and from *C* to *D* thay are at the mercy of the variable lower level winds. This results in the development of two striated layers, the upper one running south-to-north, the lower one west-to-east. Some real vision of the upper level structure is perhaps given to us by the aurora, which may be simply electrical discharges following and lighting up these ionized paths.

Such periodic structures as I have outlined could act as gigantic gratings, capable of reflecting as spectra the radiation from distant transmitters, and thus causing interference effects. For stations to our west or east, the upper layer would be more effective; for southern or northern stations the lower, because of the direction of the grating elements. According to this, there should be some difference between east and west transmission on the one hand, and north and south on the other. A com-

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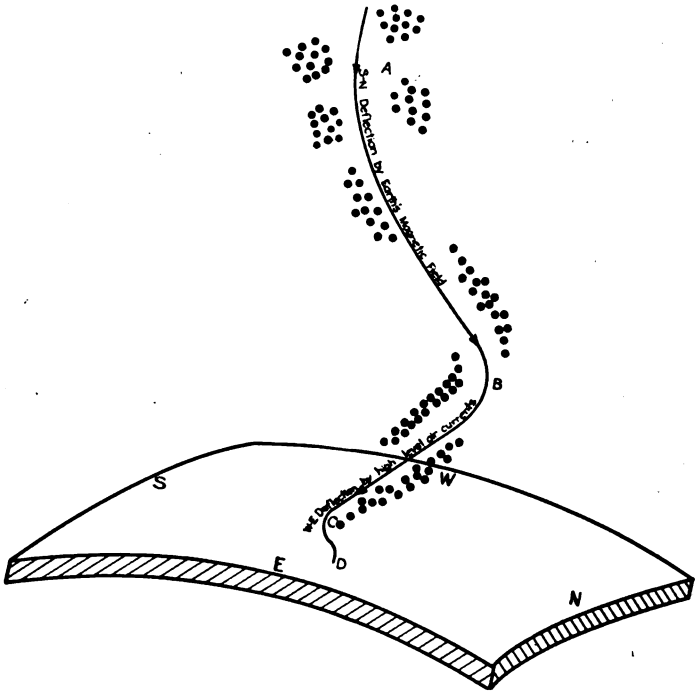

FIGURE 38

parison of a typical west-to-east transmission, such as that of Figure 27, with a typical south-to-north transmission does indeed indicate a difference in the character of the variations.

It is believed that at least a portion of the earth's magnetic field is due to these descending charges, and if this is so, we should expect to find some correspondence between magnetograph records and these transmission records. A magnetograph record does in fact show seasonal and diurnal changes, disturbances of great amplitude at times of intense aurora, and it also shows a variety of short period pulsations, measured in hours, minutes, and even seconds. Periods of 2 to 4 minutes are most frequent, while pulsations following each other at about 30 second intervals are not uncommon. Some days are relatively quiet, that is, free from rapid large amplitude oscillations, others are moderately disturbed, and many are highly disturbed. All of these effects are known in radio transmission, and some of them are shown in my records. In a later paper I hope to present simultaneous transmission and magnetograph records.

My present explanation of the transmission changes presented

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in this paper is that they are due primarily to changes in absorption, and secondarily to reflection-like effects.

Both absorption and reflection are caused by ions in our atmosphere, which I assume are not uniformly distributed, but arranged in streamer-like clouds, oriented in the upper levels by the earth's magnetic field, and in the lower atmosphere by air currents. During the day, the direct radiation from the sun adds to this ionization, producing thereby a sufficient uniformity in its distribution so that it acts principally to absorb, thereby masking any reflection effects which would be apparent in its absence. Before sunset the solar ionization decreases, and at night this additional source of ionization vanishes, and the descending ionized structure is unmasked, varying from minute to minute in its effect over a given locality as it drifts with the low level air currents. At times this structure becomes sufficiently periodic to form a two- or three-dimensional grating, and then interference effects are found, including not only transmission minima lower than any daytime value, but also changes in wavefront. Over short distances of transmission, individual or small groups of descending ions at very low levels may account for the short period swings which predominate in such transmission, so that the effect may be confined almost entirely to the lower atmosphere. Over greater distances, the effect would be a sort of statistical average of great numbers of these ion clouds, which would tend to smooth out the shorter period effects, and accentuate the longer periods corresponding with the arrival of large groups over large areas. Certainly there should be a marked correlation between transmission and weather.

Nipher[7] has found a well-defined correspondence between the horizontal intensity of the earth's magnetism and very local atmospheric disturbances. In Figure 39 his magnetograph records between 4 and 5 P. M. a sharp dip coinciding with the passage of a dense cloud. Other records show the effect of local air movement and small aread showers. Nipher's conclusion is that "local variations in the earth's magnetic field are determined wholly by local weather conditions"—perhaps the apparently very local short period variations in radio reception may be caused by changes in a comparatively minute volume of low level atmosphere.

In conclusion, I wish to acknowledge my indebtedness to Mr. H. S. Shaw, Jr., for his invaluable co-operation in this work, not

[7] Nipher, "Variations in the Earth's Magnetic Field," "Trans. Acad. of Sci. of St. Louis," volume XXII, number 4, 1913.

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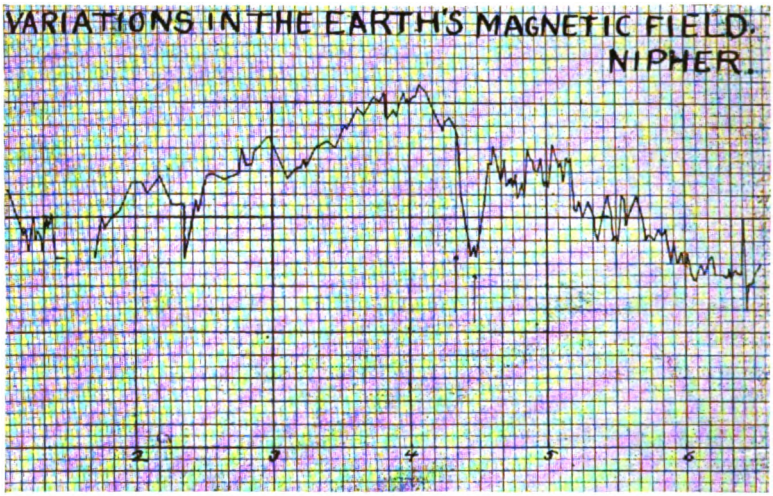

FIGURE 39

only in the design of portable apparatus for recording and in the actual taking of many of the records, but for his many helpful suggestions.

SUMMARY: Continuous records of electric field intensity at distances of from 11 to 2,600 kilometers (6.9 to 1,650 miles) from various broadcasting stations are made by amplifying and rectifying the received current from an open antenna, and recording galvanometer deflections produced by this rectified current. These records show a complex curve containing periodicities ranging from seconds to tens of minutes, the relative amplitude of the short and long period elements varying with the distance. The transition from day to night conditions is shown, and well-marked short period variations ("fading") in night-time reception from a station only 11 kilometers (6.9 miles) distant are found. Simultaneous records at separated receiving points are found to be dissimilar, even with separations as small as 550 meters (1,680 feet), and simultaneous records of different distant stations taken at the same receiving point are also shown to be entirely unlike. A tentative hypothesis for these fluctuations is advanced by the author, in which verying local absorption plays the principal part, but with marked interference effects produced by plural path transmission.

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NEW APPLICATIONS OF THE SODION DETECTOR*

By

Harold P. Donle

(Chief Engineer, Connecticut Telephone ane Electric Company, Meriden, Connecticut)

The general principles of one form of our sodion detector, which utilizes a highly electro-positive alkali metal, such as sodium, were recently described and demonstrated before The Institute of Radio Engineers. It is believed that a description of the details of another form of the sodion tube, together with a brief discussion of its operation, uses, and characteristics, will be of interest.

Possible objections to the previously described form, from some points of view, are the necessary care in handling it on account of its liquid anode and the need of operating it in a definite position. The form now to be described utilizes a solid anode, is of higher sensitiveness and is designed for economy in the filament power required. Its underlying principles are the same as those of the other form and reference is made to my earlier paper[1] for a discussion of some points not treated herein. The type S-13 sodion tube, the applications of which are the subject of the present paper, can be operated in any position, and uses a a filament current of less than ¼ ampere.

The structure of the S-13 tube is illustrated diagrammatically in Figure 1, where *C* and *A* are the collector and anode electrodes, respectively. These elements are both of nickel. The collector electrode *C* is of the same form and general proportions as in the previously described tube (that is, similar to the letter **U** with the filament *F* lying just within the open side.) The wires supporting these electrodes are sealed thru a glass bead *B* and the entire assembly mounted within a small glass tube *S*. On the outside of this tube or shell are cemented a few turns of resistance wire *R* which are connected directly in series with the filament

*Received by the Editor, October 10, 1923. Presented before The Institute of Radio Engineers, New York, October 10, 1923.

[1] "A New Non-Interfering Detector"; presented before The Institute of Radio Engineers, December 20, 1922; published on page 97 of the Proceedings for April, 1923.

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and serve the double purpose of maintaining the proper vapor pressure within the tube and further providing the necessary potential for the collector electrode. During the process of evacuation, which is carried to the highest possible degree, a small quantity of sodium is placed within the tube to provide the necessary operating characteristics.

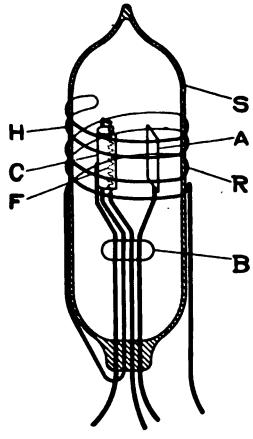

FIGURE 1

In Figure 2 all the elements of this tube are shown before assembly and also a complete tube.

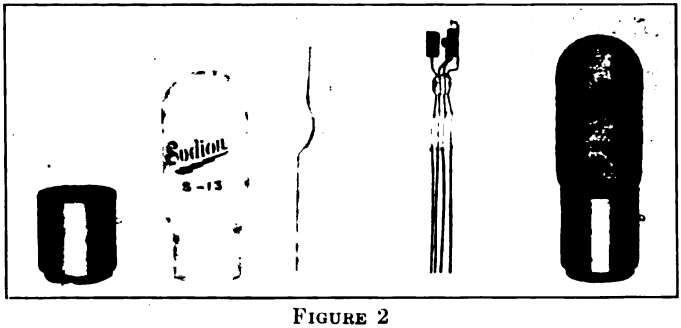

FIGURE 2

A desirable circuit for use with this tube is shown diagrammatically in Figure 3. *H* represents the heater, shown for clarity above the tube and connected in series with the negative end of the filament in order to secure a negative potential for the collector circuit. *C* and *A* are the collector and anode elec-

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trodes, and P is the potentiometer which is employed to regulate the potential of the collector for various conditions of operation. Since the greatest collector potential variation need never be more than a few tenths of a volt, there is connected in series with the positive end of the potentiometer a resistance to broaden its adjustment. This combination is connected across the tube terminals, rather than across the battery, in order that its range of adjustment will be independent of battery condition.

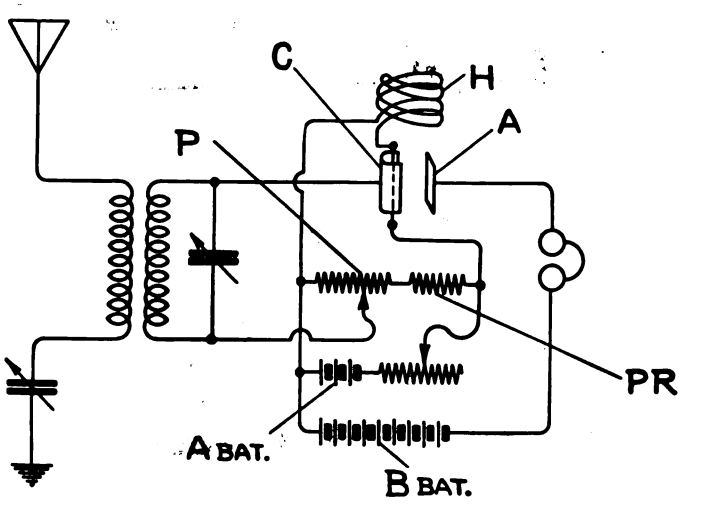

FIGURE 3

The tuning portion of Figure 3 will be recognized as the ordinary double inductively coupled circuit. The range of coupling used is considerably weaker than is common with less sensitive detectors. Thus high selectivity is secured in the antenna circuit, the exceedingly weak coupling preventing detector reaction from increasing the primary damping to an undesirable extent. The response of the sodion tube is not affected greatly by changes in coupling once the circuits are reasonably well separated, and it has been found feasible to simplify the tuner adjustments by using a fixed but weak coupling. The flexibility of the outfit is so great that this coupling need not be changed, even tho the antenna constants vary widely in different installations. Figure 4 is a receiving set using this circuit, designed particularly for use with the sodion detector and having the tuning and detector circuits in separate units. Another circuit suitable for use with this tube is shown in Figure 5. This simple arrangement has the advan-

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tage of a single tuning control, and thruout a wide frequency range gives results nearly equal to the circuit of Figure 3. A complete set of this type is shown in Figure 6.

We have found that inexperienced operators will usually secure louder signals and better selectivity with this latter circuit than with the former, due wholly to the fact that they will not properly adjust the double circuit type.

It should be noted that there is only one tuned circuit used

FIGURE 4

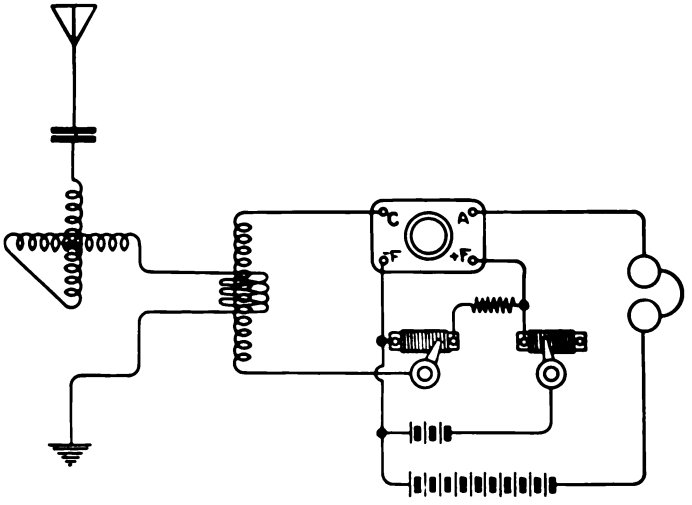

FIGURE 5

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in this arrangement, the detector input being inductively coupled. This circuit is made feasible by the high detector sensitiveness, and its inductive coupling adds materially to the selectivity.

FIGURE 6

The fundamental principles underlying the operation of this tube are complicated by a large number of factors. It is not permissable at this time to give more than an elementary description of its operation and a general sketch of some of the phenomena responsible for the reception of signals as well as of the effects produced by an alteration of the various tube adjustments. To aid in this brief explanation of the S-13 tube, the diagram shown in Figure 7 is given. This is an elementary diagram of the tube elements and external circuit, having the necessary sources of potential with an input coil *I* and indicating device *T*. Within the shell *S* is an alkali metal such as sodium, the proper conditions being maintained by the heater described above.

When the cathode or filament *F* is heated to the proper temperature by means of a battery, not shown, electrons will be evaporated from it at a certain rate. Some of these electrons will be accelerated towards the anode by the potential of the

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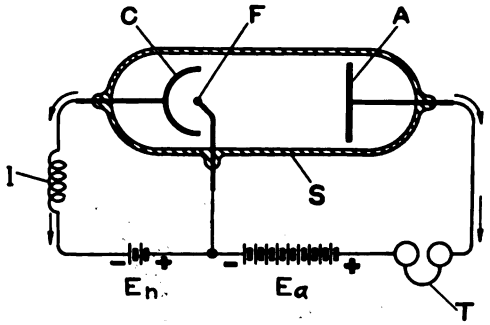

FIGURE 7

battery E_a. Some others will pass to the collector electrode C by reason of their initial velocity of emission. The number of electrons moving to this latter electrode is, however, somewhat reduced by the potential E_n, which partially neutralizes the flow of current thru the collector circuit. The density of the electronic field between filament and anode will be partly controlled by the electrostatic field between collector and filament. The number of electrons moving to the collector and, therefore, the effective potential distribution between collector and filament will be partly determined by the density of the electronic field around the filament, or, in other words, by the magnitude of the space charge.

It is convenient to consider conditions within the tube immediately after the filament is lighted. When the cathode reaches its maximum temperature and the potentials E_n and E_a are applied, there will be a flow of electrons thru the two circuits, in the direction of the arrows in Figure 7. In the anode circuit this electronic flow will be in the direction of the potential of E_a, but in the collector circuit the flow is against the potential of E_n.

So far this description applies to a pure electron tube having electrodes of the form used in the sodion. The alkali metal atmosphere of proper density which is used in these tubes, however, produces certain additional effects, a discussion of which will be interesting.

When the alkali vapor is present, some of the electrons moving between filament and anode and having the necessary ionizing velocities, will collide with atoms of the alkali metal and each such collision will result in movement toward the anode of two electrons. These two electrons are the initial electron and the one secured by ionization of the sodium atom. A positive sodium ion created by the collision will move backward into the dense

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field of electrons surrounding the filament, thereby partially neutralizing the space charge. This partial neutralization of the space charge will allow the movement of more electrons to the collector, obviously increasing the flow of current in the collector circuit, thus further overcoming the potential of the battery E_n. This change in field between filament and collector will increase the flow of electrons to A, resulting in an increase in the number of collisions and a repetition of the cycle.

From a consideration of these phenomena it is obvious that the potential distribution between filament and collector will be gradually altered and the current in the collector circuit be increased by successive small increments, each requiring an appreciable length of time on account of the relatively slow motion of the atoms concerned. These increments will cause a gradual increase in the collector current until a condition of equilibrium is attained. This condition of balance will be determined in part and may be controlled by the strength of the fields, the filament temperature, and the vapor pressure, as will be also the time required to reach equilibrium.

Under normal conditions a considerable time is required for the attainment of equilibrium on account of the mutual effects of the two circuits and the relatively slow motion of the heavy positive ions. Therefore, if a potential impulse of short duration is applied to the collector circuit, the polarity of which corresponds to a decrease in the potential of the battery E_n, the effect will be to slightly increase anode and collector currents. Since this impulse is of short duration, no time is allowed for these currents to build up by the previously described phenomena. When the applied impulse is of a polarity corresponding to an increase in the potential of the battery E_a, the initial cause of the ionization phenomena is in part removed, and this will result in a large drop of anode and collector currents, and a considerable time will be required to restore equilibrium. Consequently the increase due to a positive impulse is much less than the decrease due to a negative impulse.

When an unmodulated and relatively high radio frequency signal is applied to the coil I, complete restoration of equilibrium cannot be nearly accomplished before the arrival of the next cycle which will, of course, add to the effect produced by the first in the same direction. Thus the collector current will slowly sag by the addition of these negative potential impulses until equilibrium is re-established under this condition and at a new value determined by the amplitude of the signal impressed.

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The final result secured by this phenomenon is a large reduction of anode and collector currents when unmodulated radio frequency voltages are applied to the collector, the reduction being under complete control as to time period, and so on, and, therefore, exceedingly sensitive to weak impulses. It has even been found possible under certain conditions so to increase the time required for restoration of the balance that when spark signals were received, equilibrium was not obtained after the passage of one train before the next arrived. Therefore, it was possible to add these wave trains and to secure an extremely large drop in collector and anode currents which has distinct possibilities for telegraph calling and similar systems.

The typical static characteristics for a tube of this type, Figure 8, 9, 10, and 11 give some idea of the magnitude of these various effects. The increase of collector current due to ionization at various values of collector potential for normal operation is shown in Figure 8, together with the corresponding anode currents. It should be observed that the collector potentials given

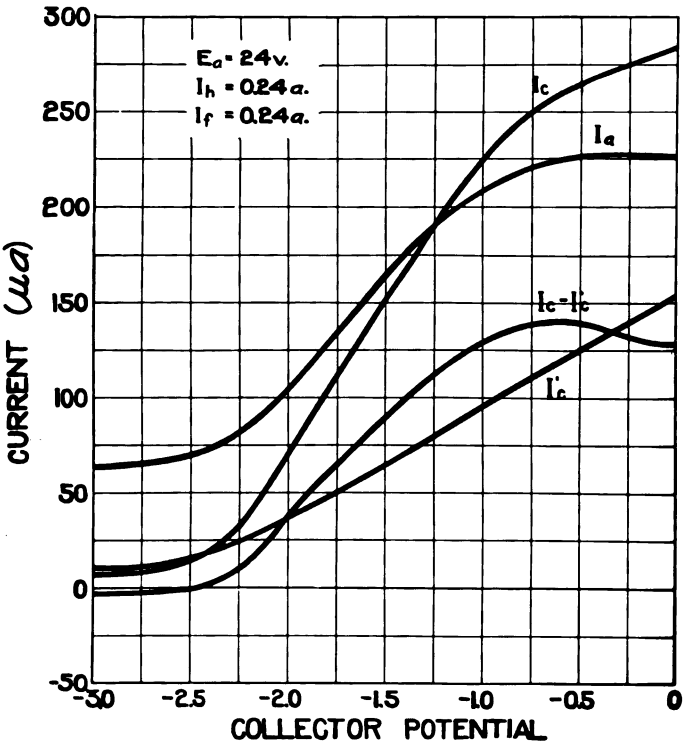

FIGURE 8

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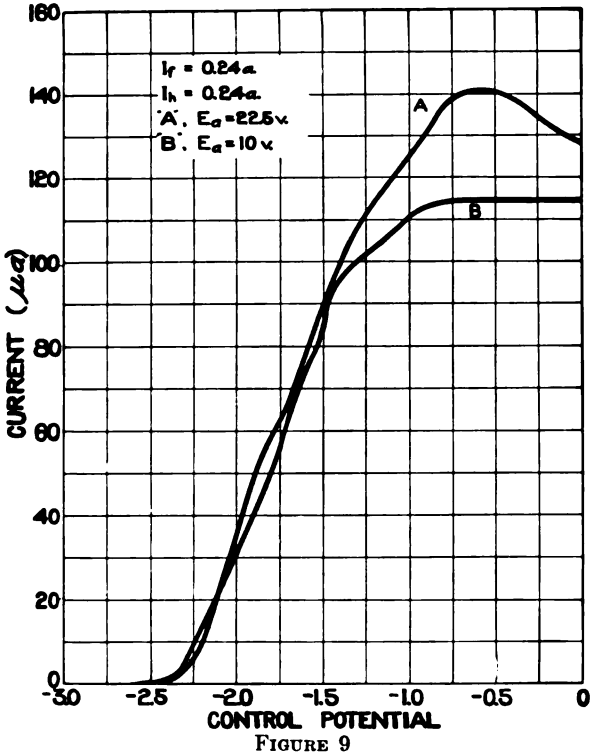

FIGURE 9

here are measured from the negative filament terminal and do not include the heater potential. The curve I_c–I_c' gives an index of the magnitude of build-up of the collector current and potential due to the above-described phenomena, I_c' being the collector current when ionization is prevented in the field between filament and anode by opening the anode circuit. Figure 9 shows the effect on this build-up of a variation of anode potential, Figure 10 the effect of a variation of heater current, and Figure 11 the variation of collector and anode currents with a variation of anode potential.

The effect of a variation of signal intensity on the change in collector current is shown in the curve of Figure 12. This curve was obtained by transmitting from one station a series of modulated signals of known relative power and observing the difference in collector current with and without the signal at a distant receiver. These readings were taken in two ways, first, with a constant value of collector potential adjusted for the maximum signal at moderate power, and secondly, with this potential

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adjusted to give the loudest signal for each of the power values used.

A consideration of the two curves of Figure 12 shows the desirability of readjusting the collector potential for various signal intensities.

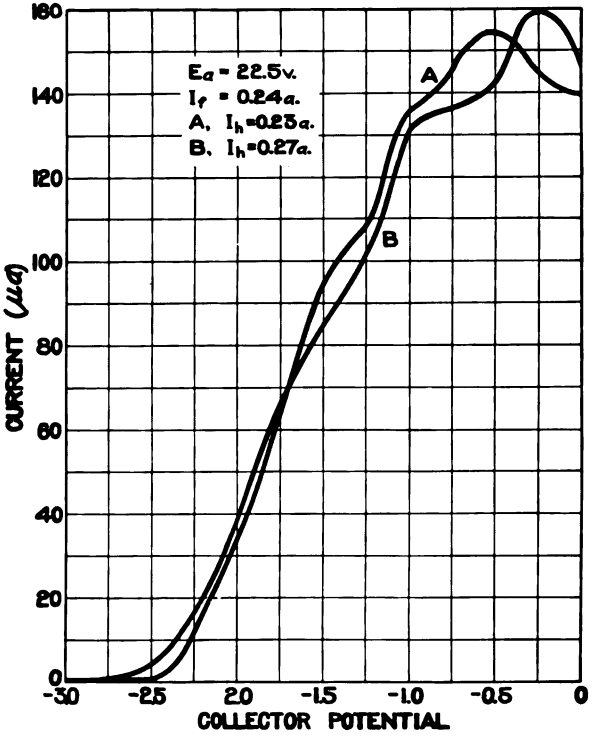

FIGURE 10

Current audibility readings on the modulated signal were also taken with telephone receivers in the anode circuit at the time data for the *B* curve of Figure 12 was secured. The results of these audibility observations are shown in Figure 13. It is interesting to note that the current audibility curve of Figure 13 and the corresponding curve *B* of Figure 12 (showing the change in collector current when adjusted for the best signal on each power) are alike in form.

Figure 14 is the relation of input radio frequency to output audio frequency. This is substantially a straight line and accounts in part for the clear undistorted signals characteristic of this detector.

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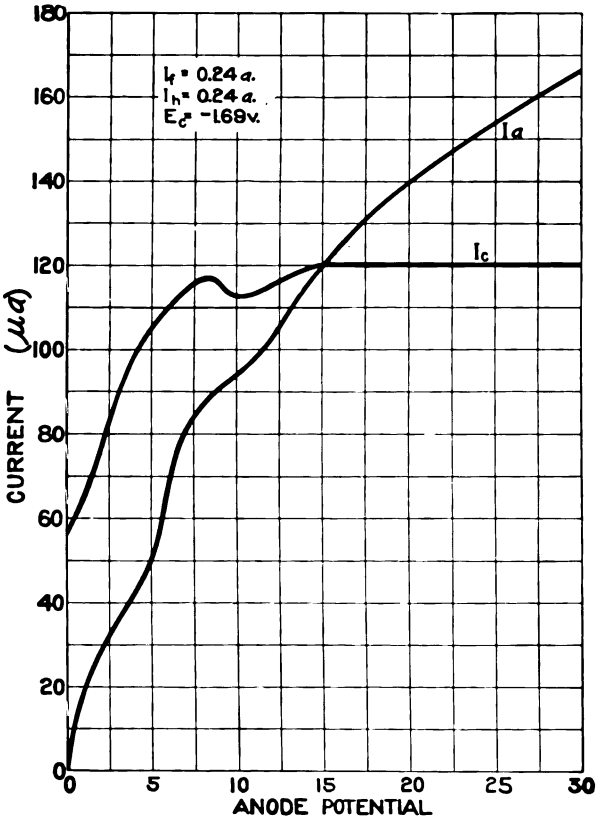

FIGURE 11

Audibility readings are of course exceedingly difficult to secure with even desirable accuracy and characteristic variations are shown by the location of points of observation shown in Figure 13. On the other hand, the micro-ammeter readings of change in collector current plotted in Figure 12 show consistency and considerable precision. It is evident from this that the sodion tube provides us with a new and accurate means of measuring the strength of weak signals.

Thus the tube has interesting possibilities for various measurement purposes, particularly on account of its simplicity and stability. For example, fading observations can be readily made with an extremely simple outfit requiring only an ordinary micro-ammeter in the collector circuit. This may have a small potential applied to its terminals thru a resistance for the purpose of reducing the current thru the meter to allow a scale of, for

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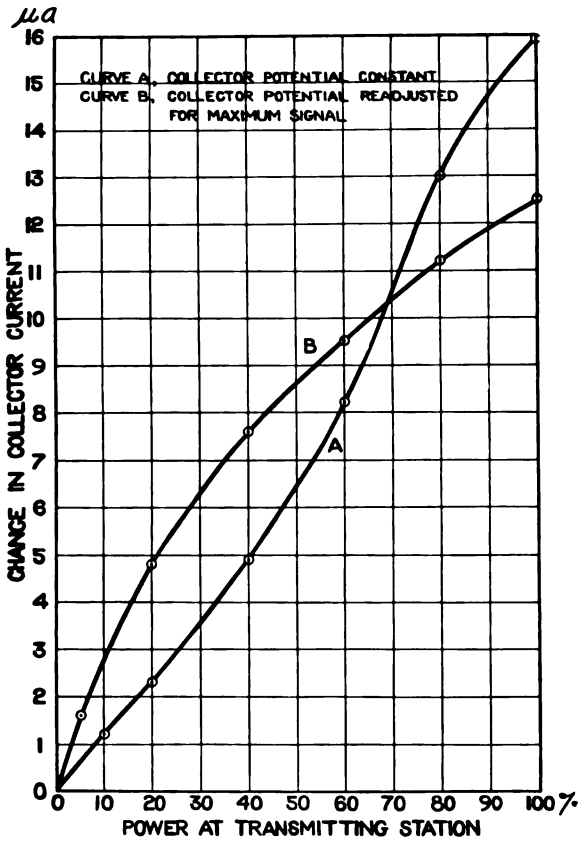

FIGURE 12

example, 100 micro-amperes to be used, and will give excellent results. With such an arrangement readings of collector current change (which have been shown to be proportional to audibility) can easily be taken every few seconds over a period of time. A typical set of observations made in this manner while receiving WGY at Meriden are given in Figure 15 and a similar set for the signal from WJZ in Figure 16.

This system can readily be elaborated by connecting several different circuits, each with its detector and micro-ammeter as described above, to the same antenna. This will allow fading observations to be taken from several different directions and distances simultaneously. Such an arrangement is perfectly practical since no interference can possibly take place between the several circuits.

The change of collector current for a signal secured by these above-described phenomena, when adjusted for telephonic re-

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ception, is very large and the energy change in the collector circuit is very nearly equal to that in the anode circuit.

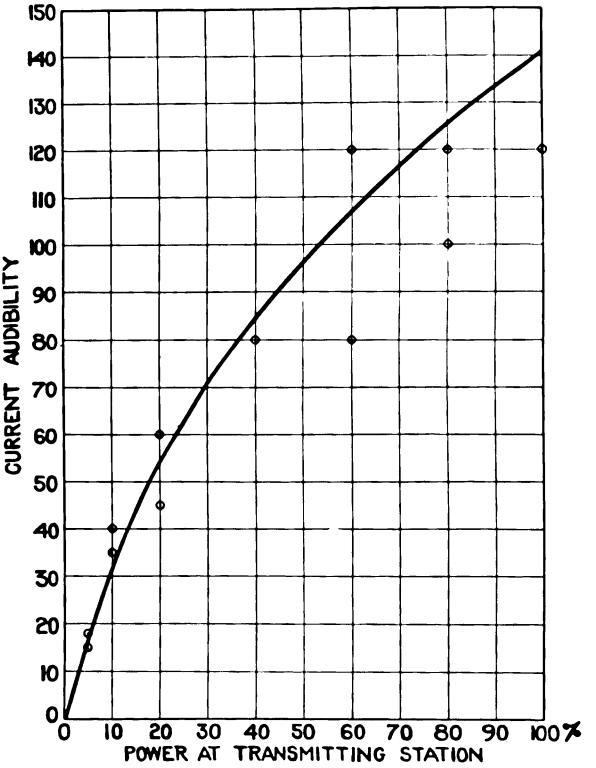

FIGURE 13

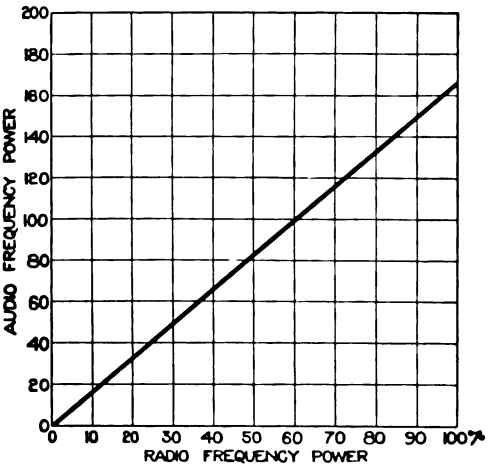

FIGURE 14

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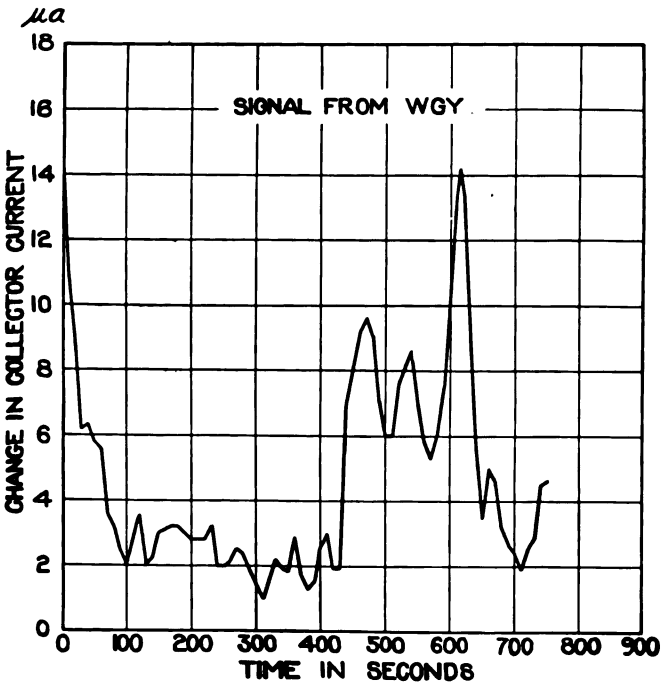

FIGURE 15

Some values of maximum collector current change for various broadcast signals are given in the table below. These were all taken under similar conditions on a small antenna in Meriden, Connecticut.

Station	*Change in collector current*
WHAM, Rochester, New York.....	5.25 micro-amperes
WEAF, New York.....	6.00 micro-amperes
WGY, Schenectady, New York.....	14.00 micro-amperes
WJZ, New York.....	4.5 micro-amperes
WDAP, Chicago, Illinois.....	5.00 micro-amperes
KDKA, Pittsburgh, Pennsylvania.....	8.50 micro-amperes

The over-all results secured with this tube are at least equal in signal magnitude to those from a regenerator or from an ordinary detector with two stages of amplification, and this without unstable adjustments, distortion, or the possibility of interference with other receiving stations. This latter point is a factor of no little importance, particularly with the rapidly increasing congestion of receivers. The tube has also a broad field in connection with radio frequency amplifying systems either with or without regeneration. The gain secured by using such a sen-

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sitive detector is the same in either case, in the former case on account of the obvious impossibility of securing the full value of regeneration more than once in the same circuit.

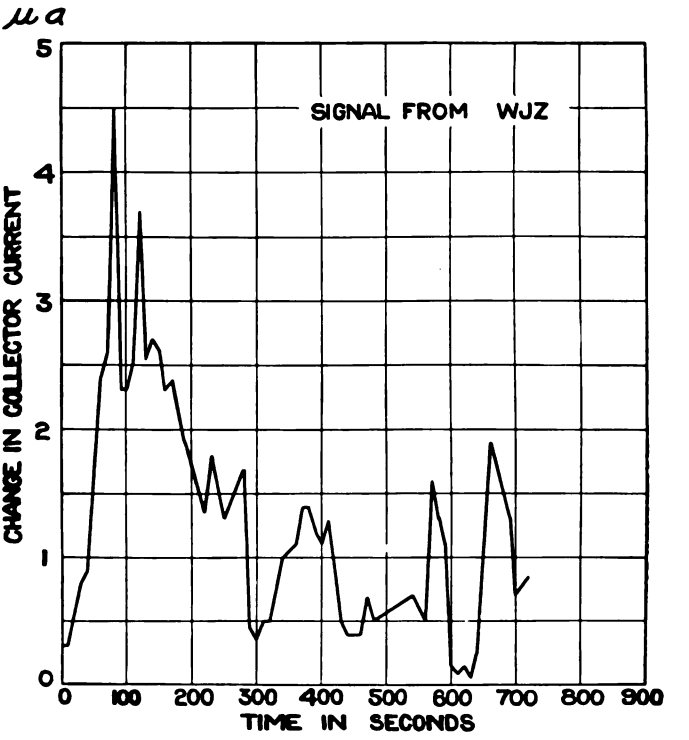

FIGURE 16

The selectivity obtainable with the circuit of Figure 5 is not equal to that of Figure 3, but the addition of a simple radio frequency filter to the circuit will allow a substantial increase in selectivity. This filter may consist of a capacity inductance circuit closely coupled to a coil of a few turns, which latter is connected in series with the ground lead from the set. A filter of this type is shown in Figure 17. The addition of such a device will make the single control circuit very selective and yet the filter adjustments do not noticeably affect the tuning of the receiving set itself.

The meter to measure change in collector current may be applied to a determination of the effectiveness of such a filter in removing a signal from the receiver. The results of a test of this kind are shown in Figure 18.

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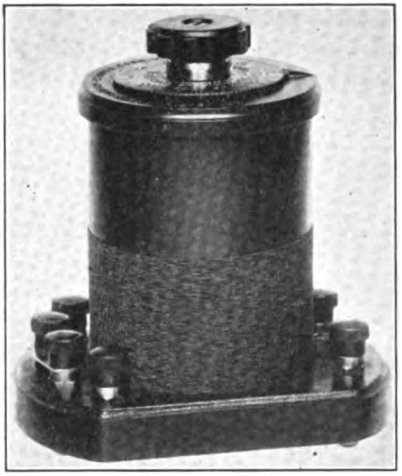

FIGURE 17

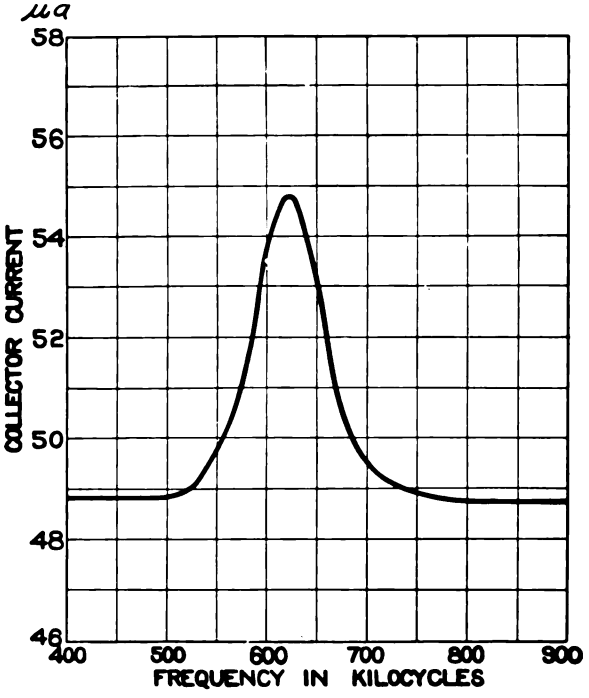

FIGURE 18

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Many other uses of the sodion tube are being investigated, and it is evident that the new device will be of utility not only as a radio receiver but as a laboratory metrical instrument.

Research Laboratory of the Connecticut
Telephone and Electric Company,
Meriden, Connecticut,
October 10, 1923.

SUMMARY: Further improvements in a new type of simple detector are described and its operation shown to be the result of ionization of sodium vapor, consequent dissipation of the space charge and the increased current to the "collector" electrode secured thereby. A condition of equilibrium is obtained and is broken up by an impressed signal. On account of the inherent and controllable time element a gradual decrease in anode current for a series of input impulses is secured, resulting in a large alteration of current for impulses of low amplitude and consequently a strong response for weak signals. Power characteristics are given and possibilities indicated of the use of this device in various types of receiving circuit and also in measurement of signal strength, such as for the observations of fading

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THE CHARACTERISTIC SURFACE OF THE TRIODE*

By

J. R. Tolmie

(Seattle, Washington)

Introduction

The purpose of this article is to outline a method of representing triode characteristics by means of surfaces rather than by curves. While at first glance this might appear to be an added complication, it is believed that it leads to a clearer physical representation than is possible by the more common method.

The current in the plate circuit of a triode is a function of the filament temperature and of the grid and plate voltages, or $I=f(T, E_g, E_p)$. This function is too complex for ordinary geometrical representation and must be simplified before a visual image of it may be obtained. This may be accomplished by assigning constant values to the variable T which make I a function only of the grid and plate voltages $I=\phi(E_g, E_p)$. Since the value of I depends simply upon the variables E_g and E_p, it may be represented as a characteristic surface, there being one and only one of these surfaces for each particular value of T. Likewise the static characteristic curves, giving the relationship between the plate current and grid voltage, are obtained by holding the filament temperature constant and by assigning fixed values to the plate voltage. Figures 1 and 2 show the two groups of static characteristic curves depending upon the grid and plate voltages, respectively; by comparing these with the characteristic surface in Figure 3 it will be seen that the curves are simply the right angle cross sections of the latter. The surface thus shows the combined action of both the grid and plate voltages in their proper geometrical relationship.

Surface Characteristics

As shown in Figure 3, the characteristic surface rises precipitously from the reference plane of zero plate current to a parallel plane corresponding to the saturation current. The

* Received by the Editor, August 21, 1923.

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directions of the co-ordinate axes are so chosen that the plate current is positive upward from the reference plane, the grid voltage is positive toward the right, and the plate voltage is

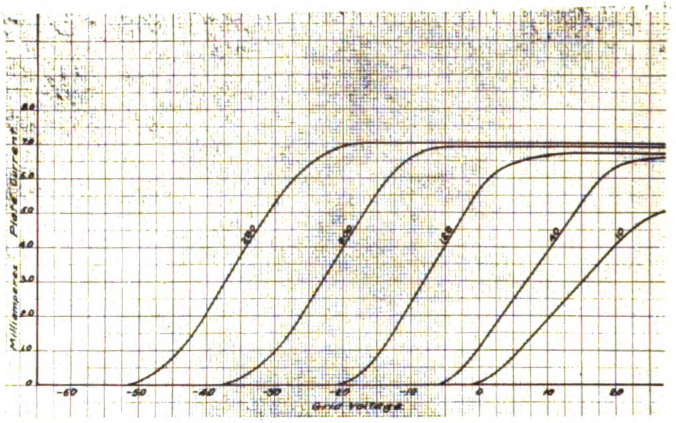

FIGURE 1

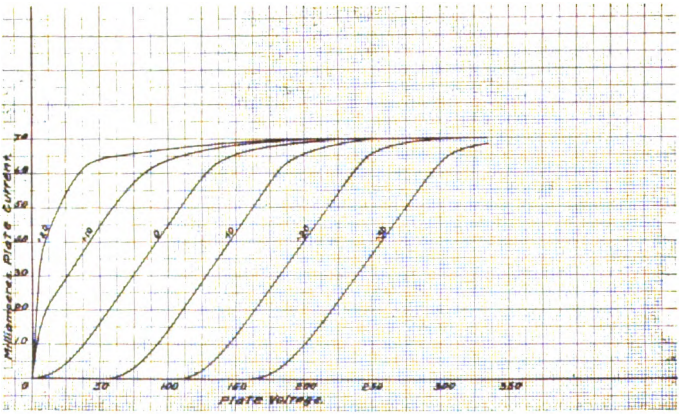

FIGURE 2

positive away from and normal to the grid voltage axis. The active portion of the surface is practically restricted to the cliff-like region running backward and toward the left from the origin or co-ordinates. It is this region that is used in amplifier and oscillator action and the upper and lower bends of which may be used for detector and modulator action. The active region extends indefinitely backward and is limited only by the permissible voltage that may be applied to the plate of the triade. As

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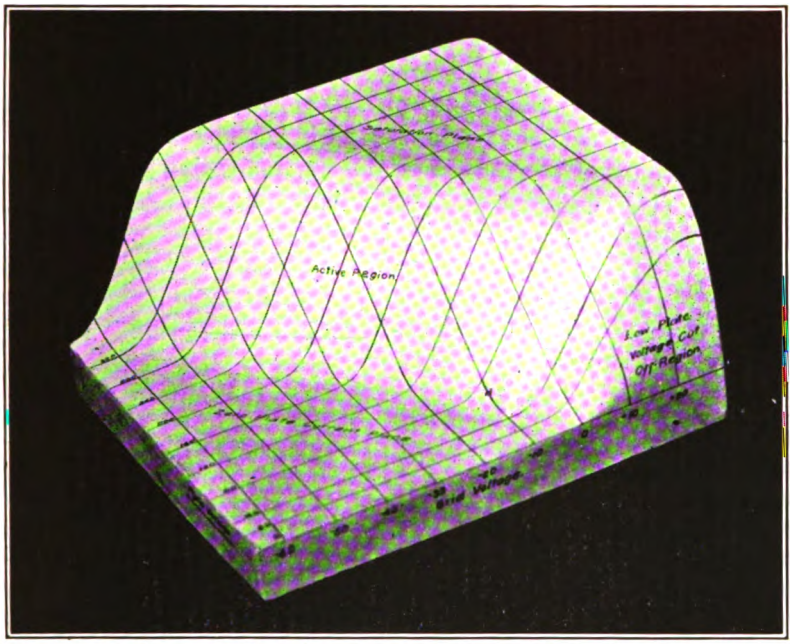

Figure 3—Characteristic surface showing Active and Cut-off Regions. Saturation current $I_s = 7$ milli-amperes

very low values of plate voltage are approached, this region swings abruptly toward the right and becomes parallel to the axis of grid voltages. This portion is practically plane and serves as a cut-off region similarly to the saturation and zero current planes. In other words, this low plate voltage region serves to limit the possible variation in plate voltage in the same way that the saturation plane serves to limit the possible variation in plate current. The right hand portion of the characteristic surface is arbitrarily bounded by a vertical plane corresponding to a relatively large value of positive grid voltage.

For high positive values of grid potential, either of two entirely different effects may manifest themselves, depending upon the internal structure of the triode. The first of these may be termed grid absorption and is due to the deflection of electrons from the plate into the grid; this is especially noticeable in triodes having fine meshed grids. The second effect occurs when electrons emitted from the filament reach sufficiently high velocities to cause secondary emission from the plate. This gives rise to a very distinctive surface lying far to the right of the one under consideration. Secondary emission is usually found in triodes

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having very coarse grids and which operate with high plate potentials.

It only remains now to note the effect of a variation in filament temperature upon the shape of the plate current surface. Any change in filament temperature will produce a like change in the number of electrons available for maintaining the plate current. This, however, will make little difference except near the saturation region, which will be either raised or lowered, depending upon whether the filament temperature has been increased or decreased. A complete set of characteristic surfaces may thus be mapped out corresponding to all the possible values of filament temperature.

The Ideal Surface

Thus far the discussion has dealt entirely with the actual characteristic surface of the triode. It will, however, simplify matters to consider a new surface toward which the actual one may approach more or less closely. Such a surface is shown in Figure 4. It will be noticed that the active region is now restricted to a skewed plane which extends clear to the axis of grid voltages. In fact, the low plate voltage cut-off region now becomes a zero-voltage cut-off plane.

Deviation from the ideal surface is caused by a number of factors, among which may be mentioned the drop in potential along the filament, the non-uniform temperature of the filament due to end supports and the limitation of plate current due to thermionic emission.

An approximate equation for the variable plate current may be obtained by expanding the functional relationship $I=f(T, E_g, E_p)$ in a Taylor's series, the expansion being taken about a point P_o corresponding to a given value of the variables T_o, E_{go}, E_{po} lying within the active region. Another relationship between the variables may also be obtained by separating them into steady and variable components.

Thus		where	
	$I=I_o+i$		$I_o<I_s$
	$T=T_o+t$		$T>0$
	$E_g=E_{go}+e_g$		$E_{po}>0$
	$E_p=E_{po}+e_p$		$E_{go}<E_{gc}$

where I_s is the saturation current and E_{gc} is a positive value of grid voltage dependent upon the shape of the low plate voltage cut-off region. For working conditions the variable temperature t is equal to zero, the fixed current I_o is equal to the normal plate current corresponding to the point P_o, e_g is a variable voltage

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impressed upon the grid, and e_p is the corresponding variation in the plate voltage. It is important to note that this latter is equal to the impedance drop with sign reversed of the current in the plate circuit; that is, $e_p = -iZ$, where z is the external plate impedance.

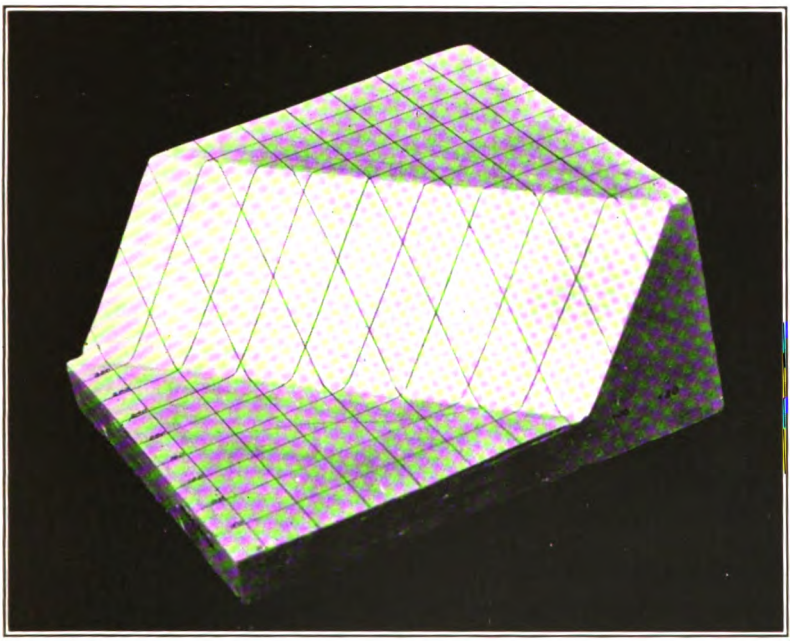

FIGURE 4—Ideal Surface. Saturation current $I_s = 7$ milli-amperes

Expanding,

$$I = I_o + \left(\frac{\partial I}{\partial E_{go}} e_g + \frac{\partial I}{\partial E_{po}} e_p\right) + \frac{1}{2!}\left(\frac{\partial^2 I}{\partial E_{go}^2} e_g^2 + 2\frac{\partial^2 I}{\partial E_{go}\partial E_{po}} e_g e_p + \frac{\partial^2 I}{\partial E_{po}^2} e_p^2\right) + \ldots \text{to } n \text{ terms.}$$

Normally this series converges so rapidly that all higher order terms may be neglected, while with an ideal surface only the first two terms are necessary.

Subtracting I_o from both sides of equation (2) and considering an ideal surface

$$i = G_m e_g + G_p e_p$$

where $G_m = \dfrac{\partial I}{\partial E_{go}}$ equals the slope of the active plane with respect to the grid voltage.

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and $G_p = \frac{\partial I}{\partial E_{po}}$ equals the slope of the active plane with respect to the plate voltage.

The quantities G_m and G_p are of the nature of conductances and are usually spoken of as the mutual and plate conductances of the triode. The reciprocal of the plate conductance $\frac{1}{G_p} = R_p$ is termed the plate resistance, while the quantity $\frac{G_m}{G_p}$ is called the amplification constant, usually denoted by μ. As may be seen by referring to Figure 4, the ratio of the two slopes or of the mutual to the plate conductance, is a measure of the relative effectiveness of a change in the grid or plate voltages in producing a given vatiation in the plate current.

Equation (3) may be transformed as follows:

$$\begin{aligned} i &= G_m e_g - G_p i Z \\ &= \frac{\mu e_g}{R_p + Z} \end{aligned} \tag{4}$$

In general $Z = R + jX$

So that $i = \frac{\mu E_{g\,max} \sin(\omega t - \theta)}{\sqrt{(R_p + R)^2 + X^2}}$

where $\theta = tan^{-1} \frac{X}{R_p + R}$ (5)

Equation (5) expresses the fact that the alternating plate current is produced by the action of an equivalent voltage of value $\mu E_{g\,max} \sin \omega t$ working into the total impedance of the plate circuit. A slightly different method of describing the same condition is to consider that the grid voltage is impressed upon a circuit having an equivalent impedance of value

$$Z_1 = \sqrt{\left(\frac{R_p + R}{\mu}\right)^2 + \left(\frac{X}{\mu}\right)^2}$$

the power factor remaining the same in both cases.

Since the plate resistance $R_p = \frac{\partial E_{po}}{\partial I}$ is also a function of the variables E_g and E_p it may likewise be shown in the form of a surface. Such a plate resistance surface taken from an actual tube is shown in Figure 5. As will be noted, the bottom of the gap corresponds to the value of resistance in the active region of the plate current surface. The sides, which rise precipitously, should continue indefinitely upward since the value of plate resistance rapidly approaches infinity, when the zero-current and saturation planes are reached. The observed data necessary for

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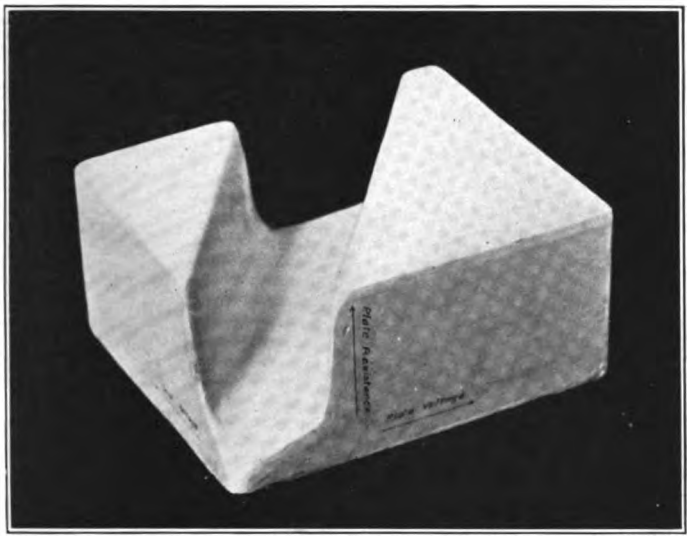

FIGURE 5—Plate resistance surface. Active value of plate resistance. $R_p = 18{,}000$ ohn s.

constructing the surfaces shown in Figures 3 and 5 were taken from a Radiotron type UV-201 amplifying triode. This particular type of triode was selected for the work because of the very wide range of voltages that may be used without causing excessive heating in either the grid or plate structures. The main considerations apply equally well, however, to any type of tungsten filament triode irrespective of its particular power rating.

OPERATION (AMPLIFICATION)

Nearly all of the properties of the triode may be described in terms of the motion of the representative point P upon the characteristic surface. This motion is not restricted to the static characteristic curves, but is perfectly general, thus allowing P to range over any point upon the surface. This gives rise to a new set of curves which are termed the dynamic characteristics. A dynamic characteristic is always obtained when a triode works into some form of a load circuit and is produced by the action of the impedance drop in that circuit. In Figure 6 is shown a set of dynamic curves corresponding to various load resistances in the plate circuit of an amplifier. The group of curves $R_{(0.....4)}$ all pass thru the operating point P_o ($E_{g_o} = -16.5$, $E_{p_s} = 160$) and, as will be noticed, decrease in slope with an increase in load resistance. This is to be expected since the greater the value of R

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the more marked will be the effect of the plate Ri drop until when $R = \infty$ there is no variation of plate current whatever and the dynamic curve becomes a horizontal line. This sheaf of dynamic curves is bounded by the horizontal curve for infinite plate resistance and by the static characteristic, shown in full line, corresponding to a load resistance of zero value. The slope of these curves depends simply upon the value of the load impedance and is not dependent to any extent upon the position of the operating point. In other words a change in the position of the operating point will simply shift the dynamic characteristic parallel to itself; see, for instance, the curves R_5 and R_6 in Figure 6. Over the greatest portion of their range the dynamic curves are very nearly straight lines and become curved only as they approach the cut-off regions.

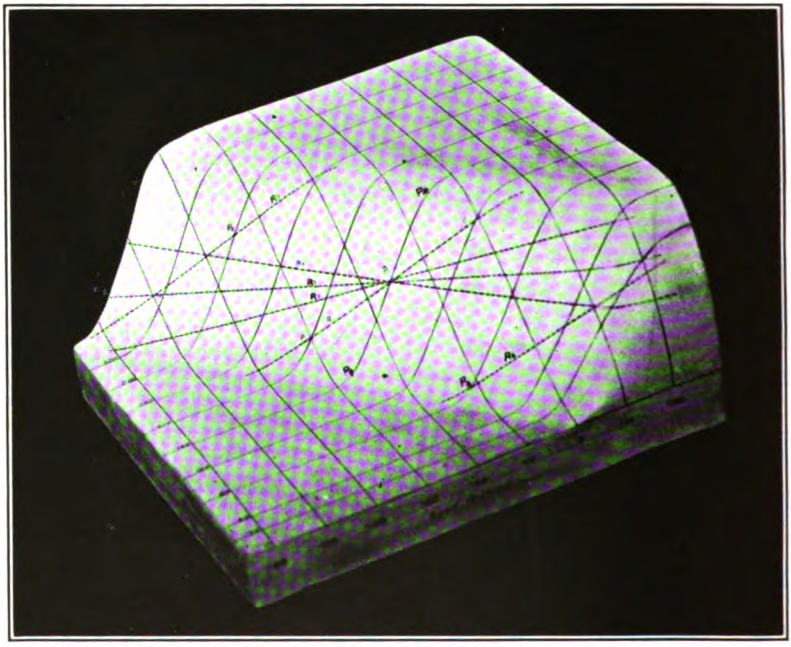

FIGURE 6—Unity power factor dynamic characteristics. Internal plate resistance = 18,000 ohms. External load resistance $R_1 = 20{,}000$ ohms, $R_2 = 50{,}000$ ohms, $R_3 = 100{,}000$ ohms, $R_4 = \infty$, $R_5 = 20{,}000$ ohms, $R_6 = 20{,}000$ ohms.

The condition for distortionless amplification is that a linear proportionality be maintained between the values of plate current and grid voltage. This means that the operating point must be kept well within the active plane of the characteristic

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surface. When this is not the case and it is allowed to approach too closely to the saturation region, as for example P_4 in Figure 6, considerable distortion will ensue. If now, however, the filament temperature be sufficiently increased, the active plane will be extended upward and correct operation may be obtained. This explains why the triode is often so sensitive to variations in filament temperature.

The triode when functioning as an amplifier is capable of amplifying either voltage, current or power; the adjustments for the three conditions are, however, quite different. To obtain a large voltage amplification it is necessary to have the dynamic, curve cut off by the low plate voltage region. This means that a large value of load resistance is required. If this is not the case, a portion of the plate voltage is not being utilized. A glance at the curves in Figure 6 will make this clear. A convenient method of obtaining the effective voltage amplification corresponding to a given load resistance is to determine the variation in plate voltage caused by a given impressed grid voltage. The ratio of these two voltages is then the quantity desired. For example, in Figure 6 with curve R_1, (A, B) a variation of 27.5 volts in the grid circuit will produce a change of 80 volts in the plate circuit; the effective amplification constant is then equal to 2.9. With an infinite load resistance the effective constant approaches its theoretical maximum, as may be easily checked from the figures, the result being 5.3, which compares favorably with the measured value of 5.5.

For current amplification, just the opposite conditions prevail. The dynamic curve should be cut off by the saturation plane and the maximum current amplification will be obtained when the load impedance is equal to zero.

The condition for best power amplification will then lie between the two conditions already outlined, or when the product of the alternating plate current and plate voltage, for a given grid nput, is a maximum. For a given value of alternating grid voltage e_g, this product is equal to

$$\frac{\mu^2 e_g^2 R}{(R_p+R)^2}$$

which is a maximum when $R=R_p$ or when the external resistance is equal to the dynamic plate resistance. This condition is shown by the curve R_1 of Figure 6, where external resistance is approximately equal to the internal resistance. The position of the operating point P is immaterial provided that the grid voltage

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s of such a value as to lie entirely within the active region of the surface.

In order to obtain maximum power output, still more rigid conditions must be specified. Stated explicitly, these are as follows: The operating point P should lie on the locus determined by $\frac{I_s}{2}$ and at as high a plate voltage as is consistent with the safe operating temperature of the triode. The upper end of the dynamic characteristic should be terminated by the intersection of the saturation plane and the low plate voltage region, while the lower end should be terminated by the zero plate current plane. These conditions allow for the maximum possible variation both in plate current and plate voltage and consequently give the maximum possible power output. Such a characteristic is shown by the curve R_2 in Figure 6. It is of course necessary to have sufficient grid voltage to work the characteristic over the entire range specified. Since the slope of the dynamic characteristic is determined by the value of the load resistance, there is, then, a particular value of load resistance for each initial position of the operating point P. In other words, there are two conditions that must be met in order to obtain maximum power output from an amplifier. *First, the load resistance must be of such a value as to produce the required dynamic characteristic corresponding to the initial position of the operating point, and second, the grid voltage must be of such a value as to cover the entire range of the above characteristic.*

It will be noticed that the curve R_5 of Figure 6 at no time passes into the region of positive grid voltages; hence, there is no tendency for electrons to flow thru the grid circuit. The only chance for power dissipation is then thru the series capacity circuits of the triode elements and load circuit or by direct leakage. The resultant power amplification under these conditions is, therefore, extremely large. Unfortunately, however, the efficiency falls to low values due to the large amount of plate voltage that remains unvaried.

The dynamic curves as shown in Figure 6 are true only for the condition of unity power factor. Whenever the load impedance is other than a true resistance, the dynamic curve takes on the form of a closed loop, such as is shown in Figure 7. These loops are due to the time phase displacement between the current and voltage waves and are accompanied neither by distortion of wave form nor dissipation of energy other than that due to the load resistance. As will be noted, the two loops are traced

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in opposite directions; the direction of motion depending upon whether the current is leading or lagging the voltage. The corresponding curve for unity power factor, where the current and voltage are in phase, is shown by Z_1.

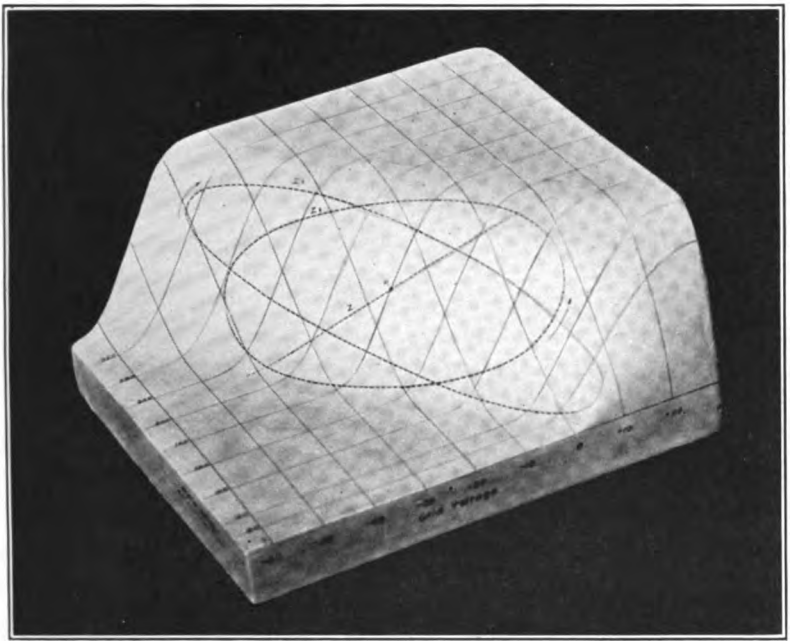

FIGURE 7—Dynamic characteristics. Internal plate resistance $R_p = 18{,}000$ ohms, Total plate impedance $Z = 38{,}000$, P.F. = 100 percent; $Z_2 = 38{,}000$ ohms. P.F. = 50 percent lagging; $Z_3 = 38{,}000$ ohms, P.F. = 50 percent leading

REGENERATIVE AMPLIFICATION

Regenerative amplification depends upon the cumulative action of the triode and its associated circuits. For instance, if an initial voltage e is impressed upon the grid circuit of a triode, a certain percentage ae of this voltage will be re-impressed upon the grid after a unit interval of time. The quantity (a) may be termed the regeneration constant and is determined by the constants of the triode and its connected circuits. The resultant voltage acting after t units of time, will then be as follows:

$$e_o = e + a\,e + a\,(e + a\,e) + a\,[e + a\,e + a\,(e + a\,e)] + \ldots\ldots \text{to } t \text{ terms} \qquad (6)$$

Expanding and collecting terms in (6)

$$e_o = e\,(1+a)^t \qquad (7)$$

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and $G_p = \frac{\partial I}{\partial E_{po}}$ equals the slope of the active plane with respect to the plate voltage.

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So that $i = \frac{\mu E_{g\,max} \sin(\omega t - \theta)}{\sqrt{(R_p + R)^2 + X^2}}$

where $\theta = \tan^{-1} \frac{X}{R_p + R}$ (5)

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$$Z_1 = \sqrt{\left(\frac{R_p + R}{\mu}\right)^2 + \left(\frac{X}{\mu}\right)^2}$$

the power factor remaining the same in both cases.

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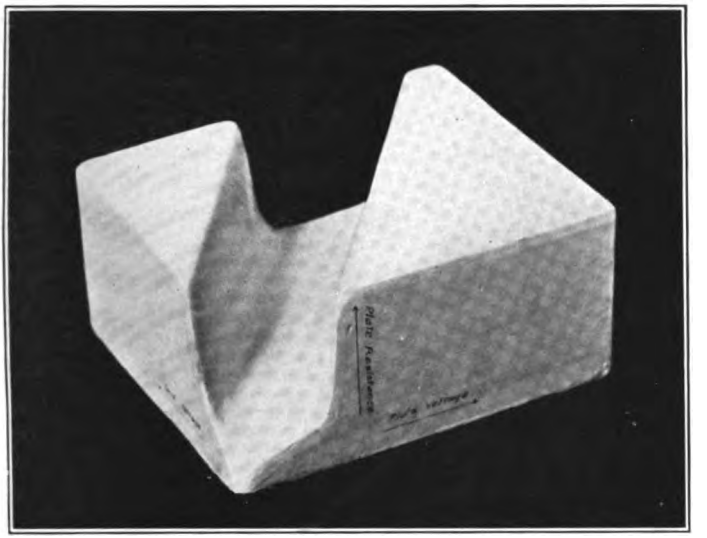

FIGURE 5—Plate resistance surface. Active value of plate resistance. $R_p = 18{,}000$ ohms.

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Nearly all of the properties of the triode may be described in terms of the motion of the representative point P upon the characteristic surface. This motion is not restricted to the static characteristic curves, but is perfectly general, thus allowing P to range over any point upon the surface. This gives rise to a new set of curves which are termed the dynamic characteristics. A dynamic characteristic is always obtained when a triode works into some form of a load circuit and is produced by the action of the impedance drop in that circuit. In Figure 6 is shown a set of dynamic curves corresponding to various load resistances in the plate circuit of an amplifier. The group of curves $R_{(0.....4)}$ all pass thru the operating point P_o ($E_{go} = -16.5$, $E_{ps} = 160$) and, as will be noticed, decrease in slope with an increase in load resistance. This is to be expected since the greater the value of R

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The resultant grid voltage thus obeys the compound interest law.

If now the voltage is considered to be reflected back m times as fast as before, the percent returned will be $\frac{a}{m}$ times that impressed and in the same interval of time there will be m times as many reflections.

Thus (7) becomes

$$e_o = e\left(1+\frac{a}{m}\right)^{mt}$$

However, $\underset{m \to \infty}{Lim}\left(1+\frac{a}{m}\right)^{m} = \varepsilon^{a}$, where ε is the base of natural logarithms.

So that $$e_o = \varepsilon^{at} e \tag{8}$$

This is the expression for the resultant voltage when the reflections become instantaneous.

The impressed voltage e may be due either to a natural oscillation in the triode circuit or to a forced oscillation from some external source. If the former, it will be of the form $e = E\,\varepsilon^{-\alpha t} \sin \omega t$ where α is the damping factor of the local circuit. The resultant grid voltage will then be

$$e_o = E\,\varepsilon^{(-\alpha+a)t} \sin \omega t \tag{9}$$

Amplification is thus produced by a decrease in the equivalent decrement of the circuit. The ultimate limit of regenerative amplification is thus reached when the equivalent decrement of the circuit has been reduced to zero; any attempt to pass beyond this point simply results in the production of local oscillation. If, on the other hand, a forced oscillation of the form $e = E \sin \omega t$ is impressed, the resulting voltage becomes

$$e_o = E\,\varepsilon^{at} \sin \omega t \tag{10}$$

As seen in equation (10) the resulting voltage tends to increase indefinitely with the time. It is, however, limited by the cut-off regions of the characteristic surface. Such an oscillation is shown by the dynamic curve of Figure 8, where the initial amplitude is determined by the intercept (A, B), the starting transient having been neglected.

In the more recently developed super-regenerative system of amplification the regeneration constant a takes on the form of a harmonic function such for example as $a = a_1 + a_2 \sin \omega_o t$, where a_1 determines the amount of fixed regeneration obtainable with a given position of the operating point and a_2 is the amplitude of the variable component corresponding to a frequency of variation

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equal to $\frac{\omega_o}{2\pi}$. This variable component is produced by moving the operating point P from the vicinity of one of the cut-off planes to a position nearer the center of the active region. For example, from P_o to P_1 in Figure 8; as has been shown by Armstring,[1] this result may be obtained either by varying the grid voltage or the plate voltage, or by varying both simultaneously. The resulting grid voltage is then expressed by

$$e_o = E\,\varepsilon^{[a_1+a_2 \sin \omega_o' t]t} \sin \omega t.$$

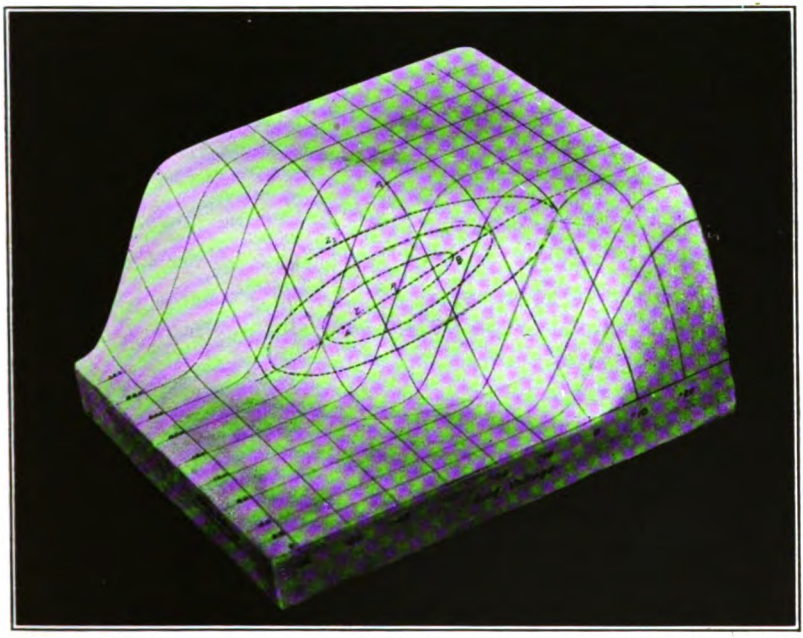

FIGURE 8—Dynamic characteristics. Internal plate resistance $R_o = 18{,}000$ ohms. Total plate impedance $Z_t = 38{,}000$ ohms, P.F. = 100 percent; $Z_g = 38{,}000$ ohms, P.F. = 87 percent lagging. Negative decrement = 1.62

This, moreover, is the same principle that is applied in constant current modulation, the two applications differing mainly in the magnitude of the control frequency corresponding to $\frac{\omega_o}{2\pi}$.

The above examples are but a few of the many possible cases where the concept of the characteristic surface may be of assistance.

[1] "Some Recent Developments of Regenerative Circuits," E. H. Armstrong, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, August, 1922.

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In conclusion the writer wishes to acknowledge his indebtedness to Mr. A. Kalin, of the Electrical Engineering Department, for the photographic reproductions; to Mr. C. J. Albrecht for many valuable suggestions on the construction of the models; and to Mr. T. M. Libby for his very kind assistance in checking the paper.

Electrical Engineering Laboratories,
University of Washington,
June 6, 1923.

SUMMARY: The paper outlines a method of representing triode characteristics by means of surfaces rather than by families of characteristic curves. The physical features of the surfaces are then described in considerable detail and the fundamental equation for the alternating component of the plate current is derived for an ideal surface.

Finally the action of the triode both as a simple and as a regenerative amplifier is described with reference to its characteristic surface.

NOTATION

SYMBOL	NAME	UNIT
a	Regeneration constant	numeric
a_1	Fixed value of regeneration constant	numeric
a_2	Amplitude of variable regeneration terms	numeric
α	Damping constant	numeric
E_{go}	Direct grid voltage	volt
E_g	Instantaneous grid voltage	volt
E_{gc}	Constant satisfying boundary conditions	volt
e_g	Variable grid voltage	volt
$E_{g\,max}$	Maximum value of alternating grid voltage	volt
e	Instantaneous impressed voltage	volt
E	Maximum value of impressed voltage	volt
e_o	Resultant voltage after regeneration	volt
E_{po}	Direct plate voltage	volt
E_p	Instantaneous plate voltage	volt
e_p	Variable plate voltage	volt
ε	2.7183 (Base of natural logarithms)	numeric
G_m	Mutual conductance	milli-mho
G_p	Plate conductance	milli-mho
I_o	Direct plate current	milli-ampere
I	Instantaneous plate current	milli-ampere
i	Variable plate current	milli-ampere
I_s	Saturation current	milli-ampere

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j	Imaginary unit	$\sqrt{-1}$
m	Multiplying factor	numeric
μ	Amplification constant	numeric
π	3.1416	numeric
R_p	Dynamic plate resistance	kilo ohm
R	External load resistance	kilo ohm
T_o	Fixed temperature	degree abs.
T	Instantaneous temperature	degree abs.
t	Variable temperature	degree abs.
t	Time	second
θ	Phase angle	radian
X	Reactance	kilo ohm
Z	Symbolic impedance	complex quantity
Z_1	Resultant impedance	kilo ohm
ω	Angular velocity	radians per second
ω_o	Modulating angular velocity	radians per second

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FORMULAS AND TABLES FOR THE CALCULATION AND DESIGN OF SINGLE-LAYER COILS*

By

Frederick W. Grover

(Union College, Schenectady, New York)

1. Introduction

Probably no form of inductance coil is of greater fundamental importance than the single-layer helical coil. Formulas are available for the calculation of the inductance of such coils which on the score of accuracy leave little to be desired. The related problem of the design of single-layer coils, that is, the predetermination of the dimensions of a coil to give a chosen inductance has received less attention. However, in view of the large number of radio workers who have occasion to construct coils for themselves, means for making such calculations with a minimum expenditure of time and labor should find an extended use.

Such tables as are at present available[1] are constructed for two variables (diameter and length) and are thus, even when voluminous, applicable to only a limited range of coil dimensions.

In the present paper advantage is taken of the fact that for the calculation of the inductance it is necessary to tabulate constants for a single parameter (the shape ratio) only. It is, therefore, possible to cover the design problems for the whole range of possible helical single-layer coils in the single moderate size table given below. The calculation of this table is based upon the accurate inductance formula of Nagaoka.[2] The design constants in the table are tabulated for the ratio $\frac{\text{diameter}}{\text{length}}$ or its reciprocal, $\frac{\text{length}}{\text{diameter}}$, values between zero and unity being included for each. Thus all possible shapes of coils are included

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[1] Circular 74, Bureau of Standards, pages 286-292, 1918. "Radio and Model Engineering," November, 1922, The Sleeper Radio Corporation, New York.

[2] "Jour. Coll. of Science," Tokyo, 27, article 6, page 18; 1909.

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that shape ratio being used in any given vase which is less than unity. No confusion results in practice from this method of procedure, and it offers the great advantage that the table may be constructed with a constant tabular interval of shape ratio thruout the whole range. The tabular interval actually chosen is sufficiently small to give a table of reasonable size, but except for a limited range at the two ends, the differences are small enough to render interpolation practically unnecessary. In the rare cases where interpolation is difficult, the desired quantity may be obtained from the value of one of the other constants for which the differences are favorable, making use of the equation connecting the quantities in question. Each problem will require a single constant to be taken from the table and none but the simplest of arithmetical calculations, which can be performed with a slide rule. An accuracy of one-half of one percent is readily attainable. For work of less accuracy, curves drawn from the tabular values may be useful.

The important case of single-layer coils wound on polygonal forms may be made to rest upon the calculation for an equivalent helical single-layer coil, and is thus included in the scope of this paper.

2. Problems to Be Solved

The inductance of a single-layer coil depends upon four variables, coil diameter, coil length, number of turns, and the shape ratio. For purposes of design the winding pitch, or better, the number of turns per unit of coil length, is more useful than the total number of turns.

Thus five types of problem are of practical importance, namely:

(a) Given the dimensions and numbers of turns (winding density), to calculate the inductance of the coil.

(b) Given the length and winding density, to calculate the diameter which the coil must have to give a desired inductance.

(c) Given the diameter and winding density, to calculate the length of coil to give a specified inductance.

(d) Given the coil diameter and coil length, to calculate the winding density necessary to give a desired inductance.

(e) Given that a certain ratio of diameter and length is desired, to calculate the dimensions necessary to give a specified inductance, when a certain winding density is assumed.

In the following section are given the formulas and procedure for the solution of each of these problems. For convenience

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formulas are given for English units as well as metric. The inductance values are in microhenrys in every case; coil dimensions are in centimeters or inches; winding density in turns per cm. or turns per inch in the metric and English systems, respectively.

For precise values of inductance a correction for insulating space should be applied. The coil length is to be taken *not* between the end turns, but as equal to the winding pitch multiplied by the whole number of turns. That is, the inductance formula applies to a coil length which extends half the winding pitch beyond the center of the last turn at each end of the coil. If this rule is followed the correction for insulating space is usually negligible except in precise work. For completeness its value may be calculated from the formula given below.

3. Summary for Working Formulas

Symbols Used:

L = the inductance of the coil in microhenrys.

d_1 = the diameter of the coil in centimeters.

d_2 = the diameter of the coil in inches.

b_1 = the length of the coil in centimeters = $\frac{n}{n_1}$.

b_2 = the length of the coil in inches = $\frac{n}{n_2}$.

n = the total number of turns on the coil.

n_1 = the number of turns per centimeter of axial length of the coil.

n_2 = the number of turns per inch of axial length.

$r = \frac{d}{b}$ = the ratio of diameter to length.

$R = \frac{b}{d}$ = the ratio of length to the diameter = $\frac{1}{r}$.

δ_1 = diameter of the bare wire, in centimeters.

δ_2 = diameter of the bare wire, in inches.

$\frac{1}{n_1}$ = winding pitch, in centimeters.

$\frac{1}{n_2}$ = winding pitch, in inches.

K = a function of r or R, given in Table 1.

$$B = k r^2 = \frac{k}{R^2}, \qquad C = \frac{k}{r} = k R, \qquad F = k r = \frac{k}{R}.$$

The values of B, C, and F are obtained from Table 1.

G and H are constants given, respectively, in Tables 2 and 3

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to be used in formulas (3) and (4), in making corrections for space occupied by insulation.

PROBLEM A—*Given diameter, length, and winding density, to calculate the inductance.*

$$L = 0.001\ k\ d_1^2\ b_1 n_1^2 \quad \text{(metric)} \tag{1}$$

The notation "metric" here and elsewhere, signifies that b_1 and d_1 are dimensions in centimeters, n_1 is in turns per centimeter, and the inductance is given in microhenrys.

$$L = 0.00254\ k\ d_2^2\ b_2\ n_2^2 \quad \text{(English)} \tag{2}$$

The notation "English" signifies that the dimensions b_2 and d^2 are in inches, n_2 is in turns per inch, and the inductance is given in microhenrys. The factor k is to be taken from Table 1 for the given value of $r = \frac{d_1}{b_1} = \frac{d_2}{b_2}$, or $R = \frac{b_1}{d_1} = \frac{b_2}{d_2}$, according as to which is less in unity. It is to be noted that r and R are both independent of the system of measurements adopted, as long as both dimensions are expressed in the same system. The dimensions b_1 and b_2 must be taken as equal to the whole number of turns, multiplied by the pitch of the winding, that is, as $\frac{n}{n_1}$ or $\frac{n}{n_2}$, as already explained.

For greater accuracy, subtract from the value of the inductance given by (1) or (2), the correction for insulating space ΔL, given in the following formula.

$$\begin{aligned} \Delta L &= 0.00628\ n_1 b_1 d_1 (G+H) \\ \Delta L &= 0.0160\ \ n_2 b_2 d_2 (G+H) \end{aligned} \tag{3}$$

or, expressed in terms of the value of L

$$\frac{\Delta L}{L} = \frac{6.28\ (G+H)}{k\ d_1\ n_1} = \frac{6.28\ (G+H)}{k\ d_2\ n_2} \tag{4}$$

The constants G and H are given, respectively, in Tables 2 and 3. G is a function of the ratio of the bare diameter of the wire to the pitch of the winding, that is, of $\delta_1\ n_1 = \delta_2\ n_2$, while H depends on the whole number of turns alone.

EXAMPLE 1:

$d_2 = 4$ inches $\qquad n_2 = 50$ turns per inch

$b_2 = 8.5$ inches $\qquad \delta_2 = 0.015$ inch

The ratio $r = \frac{4}{8.5} = 0.4706$. From column A of Table 1, we find $k = 8.17$, and thence

$$L = 0.00817\ (2.54)\ 16\ (8.5)\ (50)^2 = 7{,}056 \text{ microhenrys.}$$

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To see if the correction for insulating space need to be considered, we find $n_2 \delta_2 = 0.75$, and thence, from Table 2, $G = 0.27$. The whole number of turns is $n = 50\ (8.5) = 425$, so that from Table 3 we find $H = 0.335$. The fractional correction is, therefore

$$\frac{\Delta L}{L} = \frac{6.28\ (0.605)}{8.17\ (200)}, \text{ or only 2.3 part in a thousand.}$$

Given diameter, length, and the total number of turns, n, to calculate the inductance

$$L = 0.001\ F\ d_1\ n^2 \text{ (metric)} \qquad (5)$$

$$L = 0.00254\ F\ d_2\ n^2 \text{ (English)} \qquad (6)$$

The ratio r or R has to be calculated (that one which is less than unity being used) and in Table 1 the value of F is taken corresponding to that ratio.

EXAMPLE 2—To calculate the inductance of a coil having 79 turns, wound on a form of diameter 22.3 cm. The axial distance between the first and the 79th wire was found to be 9.0 cm. Thus the pitch $\frac{1}{n_1} = 0.115$ cm., so that the equivalent length of the coil was 79 times $0.115 = 9.12$ cm.

$$d_1 = 22.3, \quad b_1 = 9.12, \quad n = 79, \quad R = \frac{9.12}{22.3} = 0.409.$$

From Table 1 the value of F corresponding to this value of R is 11.515, so that for the inductance we have

$$L = 0.011515\ (79)^2\ 22.3 = 1602.6.$$

Supposing the wire to have a bare diameter of 0.8 mm., $n_1 \delta_1 = 0.7$, and from Tables 2 and 3, we find $G = 0.20$, $H = 0.33$, and thus

$$\Delta L = 0.00628\ (79)\ 22.3\ (0.53) = 6.9 \text{ microhenrys}$$

so that the inductance, corrected for insulating space, is 1,597 microhenrys.

PROBLEM B—*Given inductance, length, and winding density, to find the diameter.*

Calculate from the given data, the quantity.

$$B = \frac{1000\ L}{b_1^3\ n_1^2} \text{ (metric)} \qquad (7)$$

or

$$B = \frac{393.7\ L}{b_2^3\ n_2^2} \text{ (English)} \qquad (8)$$

and find in Table 1, column B, the value of r or R corresponding to this value of B. From this value calculate the diameter from the relation $d_1 = b_1 r = \frac{b_1}{R}$, or $d_2 = b_2 r = \frac{b_2}{R}$.

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In general, for purposes of design, it will not be worth while to evaluate the corrections for the insulating space, but in case this is done, it will be first necessary to make the approximate solution, as just shown. With the value of d_1 thus found, the correction to L is found by (3), and the calculation is repeated, using the corrected value of L. This more accurate value of d_1 will not differ much from the first approximation.

EXAMPLE 3—To find the diameter of a coil which, when wound with 40 turns per inch for a length of 6 inches, will give an inductance of 5,100 microhenrys.

$$L = 5{,}100, \quad b_2 = 6, \quad n_2 = 40.$$

Therefore $B = \dfrac{393.7\,(5{,}100)}{216\,(1600)} = 5.810$, and in column B of Table 1 we find this value of B corresponds to $r = 0.912$. Thus $d_2 = 6\ (0.912) = 5.47$ inches.

PROBLEM C—*Given the diameter, and the winding density to calculate the length of coil necessary to give a specified inductance L.*

Calculate from the given data, the quantity

$$C = \frac{1000\,L}{d_1^3\,n_1^2} \quad \text{(metric)} \tag{9}$$

or

$$C = \frac{393.7\,L}{d_2^3\,n_2^2} \quad \text{(English)} \tag{10}$$

With this value of C, find in column C of Table 1, the value of r or R corresponding. Then $b_1 = \dfrac{d_1}{r} = d_1 R$, and $b_2 = \dfrac{d_2}{r} = d_2 R$. The total number of turns will be $n = n_1 b_1 = n_2 b_2$.

EXAMPLE 4—Wire is to be wound 15 turns per centimeter on a cylinder 10 centimeters in diameter to obtain an inductance of 1,000 microhenrys. To calculate the length of coil necessary we have $C = \dfrac{1{,}000(1{,}000)}{1000\,(225)} = 4.44$, and this value of C in column C of Table 1 corresponds to $R = 0.730$. Thus $b_1 = 0.730\ (10) = 7.30$ cm. The total number of turns is 109.5, so that 110 would give the required inductance.

PROBLEM D—*Given the dimensions of a winding form, to calculate the winding density necessary in order to obtain a specified inductance.*

From the given dimensions calculate the shape ratio r, or R, (less than unity) and with this ratio, find in column D of Table 1 the value of k corresponding. Then the winding density is given by

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$$n_1^2 = \frac{1000\,L}{k d_1^2 b_1} \quad \text{(metric)} \tag{11}$$

$$n_2^2 = \frac{393.7\,L}{k\,d_2^2 b_2} \quad \text{(English)} \tag{12}$$

EXAMPLE 5—How many turns must be wound per inch of axial length on a coil which is to have a diameter of 5.5 inches, a length of 9.6 inches, and which is to have an inductance of 7,100 microhenrys?

These dimensions give $r = \frac{5.5}{9.6} = 0.573$, and the value of k, corresponding to this value of r, is 7.86. Thus

$$n_2 = \sqrt{\frac{393.7\,(7100)}{(5.5)^2\,9.6\,(7.86)}} = 35.0 \text{ per inch.}$$

Wound thus, the coil will have a total of 336 turns.

PROBLEM E—*A coil is to have a given shape ratio (relation of length to diameter). When wound with a given winding density, what must the dimensions be in order that the inductance may have a specified value?*

The given shape ratio determines r and R. With the given value of r or R, obtain the value if C corresponding in column C of Table 1, and with this the diameter of the coil may be obtained from

$$d_1^3 = \frac{1000\,L}{C\,n_1^2} \quad \text{(metric)} \tag{13}$$

$$d_2^3 = \frac{393.7\,L}{C\,n_2^2} \quad \text{(English)} \tag{14}$$

Thence the length is obtained from $b_1 = \frac{d_1}{r} = d_1 R$, or $b_2 = \frac{d_2}{r} = d_2 R$.

Or we may solve for the length directly, instead of the diameter, by obtaining for the given value of r or R, the number in column B of Table 1 corresponding, and then

$$b_1^3 = \frac{1000\,L}{B\,n_1^2} \quad \text{(metric)} \tag{15}$$

$$b_2^3 = \frac{393.7\,L}{B\,n_2^2} \quad \text{(English)} \tag{16}$$

With this value of the length, the diameter is found from

$$d_1 = b_1\,r = \frac{b_1}{R}, \text{ or } d_2 = b_2\,r = \frac{b_2}{R}.$$

EXAMPLE 6—A coil is to be wound on a form for which the

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ratio of diameter to length is 2.6, and the wire is to be wound 5 turns per centimeter. What must the dimensions of the coil be, if an inductance of 1,000 microhenrys is desired?

To solve for d_1, we have $R = \frac{1}{2.6} = 0.3846$, to which the value 1.756 in column C of Table 1 corresponds. Thence $d_1^3 = \frac{(1000)^2}{1.756\,(25)}$ or $d_1 = 28.35$ cm., and therefore $b_1 = 0.3846\ (28.35) = 10.90$ cm.

The number corresponding to $R = 0.3846$ in column B of Table 1 is 30.85, which gives $b_1^3 = \frac{(1000)^2}{30.85\,(25}$), or $b_1 = 10.90$, and $d_1 = \frac{10.90}{0.3846} = 28.35$. The two solutions check one another. The coil will have a total of 54.5 turns.

4. Calculation of Polygonal Coils

Let

N = number of sides of the polygon.

s = length of side of the polygon.

D = diameter of the circumscribed circle.

d = diameter of the equivalent circular coil (same inductance.

b = axial length of the polygon coil = number of turns divided by the winding number.

Since this section deals only with the determination of the diameter of the circular coil which has the same inductance as the given polygonal coil, only ratios enter, and thus it is immaterial what units are used for the dimensions, as long as the same system is used thruout. Thus the symbols s, D, b, and d, refer to either system. When, however, the diameter of the equivalent circular coil has been found, the farther solution of the problem is to be made with the formulas for circular coils, and the formula appropriate to the system of units desired must be chosen.

If the polygon has an even number of sides, the diameter of the circumscribed circle is readily found by calipering over two opposite vertices of the polygon, and subtracting from this the diameter of cross section of the wire. If the number of sides is odd, or in other cases where the given dimension is the length of side of the polygon, the diameter of the circumscribed circle if given by

$$D = \frac{a}{\sin \frac{\pi}{N}} \qquad (17)$$

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Calculate the ratio $\rho = \frac{b}{D}$. Then from Table 4, the value of $\frac{d}{D}$ corresponding to this value of ρ is to be taken. This enables the diameter d of that circular coil to be found which, wound with the same number of turns n, and having the same length b, has the same inductance as the polygonal coil. Knowing the dimensions of the equivalent circular coil, the inductance may be calculated by (1) or (2), or (5) or (6).

For the problems of design, the calculations of Problem B, C, D, or E are to be made, using the given values of n and b, and where d enters, the value of the equivalent diameter as just described. In Problem B and E, where the diameter d is required, the value calculated by the formulas of these sections is the diameter of the equivalent coil, and the diameter of the circle circumscribed around the desired polygon has to be found. Calculating $\frac{b}{d}$, the value corresponding to the value of d already found, the value of $\frac{d}{D}$ for this value of $\frac{b}{d}$ assumed approximately equal to $\frac{b}{D}$ is taken from Table 4, and thus a first approximation to the desired value of D can be calculated. Using this a new value of $\frac{b}{D}$ is calculated and used in Table 4 to give a second and very accurate value of $\frac{d}{D}$, and thus of D. Formula (7) will then give the required length of a side of the polygonal coil. An example will make these points clearer.

EXAMPLE 7—A hexagonal coil has a side of 10 centimeters and is wound with wire 10 turns to the centimeter, for an axial length of 6 centimeters. To calculate the inductance we have $b_1 = 6$, $n_1 = 10$, $s = 10$. From (17) we find for the diameter of the circumscribed circle the value of 20 cm. (a value evident from elementary geometry also). Thus $\rho = \frac{6}{20} = 0.3$, and Table 4 gives for this value, $\frac{d}{D} = 0.912$. Thus the equivalent circular coil has a diameter of $d = 20\ (0.912) = 18.24$ cm.

The inductance is then to be calculated by formula (1). $R = \frac{6}{18.24} = 0.329$, and from Table 1, for this value of R, the value of k is 4.20, so that the inductance is

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$L = 0.001\ (4.20)\ (18.24)^2\ 6\ (10)^2 = 838$ microhenrys. Formula (5) gives, using column F in Table 1, $n = 60$,

$L = 0.001\ (12.79)\ 18.24\ (60)^2 = 840$ microhenrys.

The check is sufficiently close.

In Problem B, Case 3, the design has given $d_2 = 5.47$ inches. If we wish to find the equivalent octagonal coil having the same length $b_2 = 6$ inches, and the same number of turns, the calculation is made as follows:

$\frac{d_2}{b_2} = \frac{5.47}{6} = 0.912$, and if we assume this as a first approximation to the ratio $\frac{D}{b}$, we find from Table 4 for this value, $\frac{d}{D} = 0.949$.

Thus to a first approximation $D = \frac{5.47}{0.949} = 5.76$ inches.

Using this, we find as a second approximation to $\frac{D}{b}$, $\frac{5.76}{6} = 0.96$, and the value of $\frac{d}{D}$ corresponding to this in Table 4 does not differ from the previous value by as much as one part in a thousand. Thus we may take as final $D = 5.76$ inches, to which, by (17) the length of side is $s = 5.76 \sin 22°.5 = 2.20$ inches.

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TABLE 1

Constants for the Calculation and Design of Single-layer Coils or Solenoids as Functions of the Shape Factor

$$r = \frac{diameter}{length} \quad \text{or} \quad R = \frac{length}{diameter}$$

r	k	B	C	F	r	k	B	C	F
	Probs. A & D	Prob. B	Prob. C	Prob. A		Probs. A & D	Prob. B	Prob. C	Prob. A
0.00	9.87	0	Infinite	0	0.25	8.90	0.556	35.6	2.225
.01	9.83	0.0010	983	0.098	.26	8.86	.599	34.1	2.304
.02	9.79	.0039	489	.196	.27	8.83	.644	32.7	2.384
.03	9.745	.0088	325	.292	.28	8.79	.689	31.4	2.462
.04	9.70	.0155	243	.388	.29	8.76	.736	30.2	2.540
0.05	9.66	0.0242	193	0.483	0.30	8.72	0.785	29.1	2.617
.06	9.62	.0346	160	.577	.31	8.69	.835	28.0	2.693
.07	9.58	.0470	137	.671	.32	8.65	.886	27.0	2.769
.08	9.54	.0611	119	.763	.33	8.62	.939	26.1	2.844
.09	9.50	.0770	106	.855	.34	8.59	.993	25.2	2.919
0.10	9.46	0.0946	94.6	0.946	0.35	8.55	1.048	24.4	2.993
.11	9.42	.114	85.7	1.037	.36	8.52	1.104	23.7	3.067
.12	9.385	.135	78.2	1.126	.37	8.49	1.162	22.9	3.140
.13	9.35	.158	71.9	1.215	.38	8.45	1.221	22.2	3.212
.14	9.31	.182	66.5	1.303	.39	8.42	1.281	21.6	3.284
0.15	9.27	0.208	61.8	1.390	0.40	8.39	1.342	20.97	3.355
.16	9.23	.236	57.7	1.477	.41	8.36	1.404	20.38	3.426
.17	9.19	.266	54.1	1.563	.42	8.32	1.468	19.82	3.496
.18	9.16	.297	50.9	1.648	.43	8.29	1.533	19.28	3.565
.19	9.12	.329	48.0	1.732	.44	8.26	1.599	18.77	3.634
0.20	9.08	0.363	45.4	1.816	0.45	8.23	1.666	18.29	3.703
.21	9.04	.399	43.1	1.899	.46	8.20	1.735	17.82	3.771
.22	9.01	.436	40.9	1.982	.47	8.17	1.804	17.37	3.838
.23	8.97	.475	39.0	2.063	.48	8.14	1.874	16.95	3.905
.24	8.935	.515	37.2	2.144	.49	8.105	1.946	16.54	3.972
0.25	8.905	0.556	35.6	2.225	0.50	8.075	2.019	16.15	4.037

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TABLE 1—Continued

r	k	B	C	F	r	k	B	C	F
	Probs. A & D	Prob. B	Prob. C	Prob. A		Probs. A & D	Prob. B	Prob. C	Prob. A
0.50	8.075	2.019	16.15	4.04	0.75	7.38	4.15	9.84	5.535
.51	8.04	2.092	15.77	4.10	.76	7.35	4.25	9.68	5.59
.52	8.01	2.167	15.41	4.17	.77	7.33	4.35	9.52	5.64
.53	7.985	2.243	15.07	4.23	.78	7.30	4.44	9.365	5.70
.54	7.96	2.320	14.73	4.30	.79	7.28	4.54	9.215	5.75
0.55	7.93	2.398	14.41	4.36	0.80	7.255	4.64	9.07	5.80
.56	7.90	2.476	14.10	4.42	.81	7.23	4.74	8.93	5.86
.57	7.87	2.556	13.80	4.485	.82	7.21	4.845	8.79	5.91
.58	7.84	2.637	13.52	4.55	.83	7.18	4.95	8.65	5.96
.59	7.81	2.719	13.24	4.61	.84	7.16	5.05	8.52	6.01
0.60	7.78	2.802	12.97	4.67	0.85	7.13	5.15	8.39	6.06
.61	7.75	2.885	12.71	4.73	.86	7.11	5.26	8.27	6.115
.62	7.73	2.97	12.46	4.79	.87	7.09	5.36	8.15	6.165
.63	7.70	3.06	12.22	4.85	.88	7.06	5.47	8.03	6.22
.64	7.67	3.14	11.99	4.91	.89	7.04	5.58	7.91	6.27
0.65	7.64	3.23	11.76	4.97	0.90	7.02	5.68	7.80	6.315
.66	7.62	3.32	11.54	5.03	.91	6.99	5.79	7.69	6.365
.67	7.59	3.41	11.33	5.085	.92	6.97	5.90	7.58	6.41
.68	7.56	3.50	11.12	5.14	.93	6.95	6.01	7.47	6.46
.69	7.54	3.59	10.92	5.20	.94	6.93	6.13	7.34	6.51
0.70	7.51	3.68	10.73	5.26	0.95	6.90	6.23	7.27	6.56
.71	7.48	3.77	10.54	5.31	.96	6.88	6.34	7.17	6.61
.72	7.46	3.87	10.36	5.37	.97	6.86	6.45	7.07	6.65
.73	7.43	3.96	10.18	5.425	.98	6.84	6.57	6.98	6.70
.74	7.41	4.055	10.01	5.48	.99	6.82	6.68	6.885	6.75
0.75	7.38	4.15	9.84	5.535	1.00	6.79	6.79	6.79	6.79

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TABLE 1—Continued

R	k	B	C	F	R	k	B	C	F
	Probs. A & D	Prob. B	Prob. C	Prob. A		Probs. A & D	Prob. B	Prob. C	Prob. A
1 00	6.79	6.79	6.79	6.79	0.75	6.15	10.93	4.61	8.20
0.99	6.77	6.91	6.705	6.84	.74	6.13	11.17	4.53	8.27
.98	6.75	7.03	6.62	6.89	.73	6.09	11.42	4.44	8.34
.97	6.73	7.15	6.53	6.94	.72	6.05	11.68	4.36	8.41
.96	6.71	7.28	6.44	6.985	.71	6.02	11.94	4.275	8.48
0.95	6.68	7.405	6.35	7.035	0.70	5.99	12.22	4.19	8.55
.94	6.66	7.54	6.26	7.085	.69	5.95	12.51	4.11	8.63
.93	6.64	7.67	6.17	7.14	.68	5.92	12.80	4.03	8.705
.92	6.61	7.81	6.08	7.19	.67	5.88	13.11	3.94	8.78
.91	6.59	7.96	6.00	7.24	.66	5.85	13.43	3.86	8.86
0.90	6.56	8.10	5.91	7.29	0.65	5.81	13.76	3.78	8.94
.89	6.54	8.255	5.82	7.35	.64	5.78	14.10	3.70	9.025
.88	6.51	8.41	5.73	7.40	.63	5.74	14.46	3.615	9.11
.87	6.49	8.57	5.64	7.46	.62	5.70	14.83	3.53	9 195
.86	6.46	8.74	5.56	7.51	.61	5.66	15.22	3.45	9.28
0.85	6.44	8.91	5.47	7.57	0.60	5.62	15.62	3.37	9.37
.84	6.41	9.08	5.38	7.63	.59	5.58	16.04	3.29	9.46
.83	6.38	9.26	5.30	7.69	.58	5.54	16.47	3.21	9.55
.82	6.35	9.45	5.21	7.75	.57	5.50	16.93	3.135	9.65
.81	6.33	9.64	5.12	7.81	.56	5.46	17.40	3.06	9.75
0.80	6.30	9.84	5.04	7.87	0.55	5.415	17.90	2.98	9.845
.79	6.27	10.045	4.95	7.935	.54	5.37	18.42	2.90	9.95
.78	6.24	10 26	4.87	8.00	.53	5.33	18.96	2.82	10.05
.77	6.21	10.47	4.78	8.06	.52	5.28	19.53	2.75	10.15
.76	6.18	10.70	4.70	8.13	.51	5.23	20.12	2.67	10.26
0.75	6.15	10.93	4.61	8.20	0.50	5.19	20.74	2.59	10.37

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TABLE 1—Continued

R	k	B	C	F	R	k	B	C	F
	Probs. A & D	Prob. B	Prob. C	Prob. A		Probs. A & D	Prob. B	Prob. C	Prob. A
0.50	5.19	20.7	2.593	10.37	0.25	3.61	57.7	0.902	14.43
.49	5.14	21.4	2.518	10.48	.24	3.52	61.1	.845	14.67
.48	5.09	22.1	2.443	10.60	.23	3.43	64.9	.790	14.93
.47	5.04	22.8	2.368	10.72	.22	3.34	69.1	.736	15.20
.46	4.99	23.6	2.294	10.84	.21	3.25	73.7	.683	15.48
0.45	4.935	24.4	2.221	10.97	0.20	3.16	78.9	0.631	15.78
.44	4.88	25.2	2.148	11.095	.19	3.06	84.8	.581	16.10
.43	4.83	26.1	2.076	11.23	.18	2.96	91.3	.532	16.43
.42	4.77	27.05	2.004	11.36	.17	2.85	98.7	.485	16.78
.41	4.715	28.05	1.993	11.50	.16	2.74	107.2	.439	17.15
0.40	4.66	29.1	1.863	11.64	0.15	2.63	119.7	0.395	17.55
.39	4.60	30.2	1.793	11.79	.14	2.52	128.4	.352	17.98
.38	4.54	31.4	1.724	11.94	.13	2.40	141.8	.312	18.44
.37	4.475	32.7	1.656	12.095	.12	2.27	157.8	.273	18.93
.36	4.41	34.0	1.588	12.26	.11	2.14	177.0	.236	19.47
0.35	4.35	35.5	1.522	12.42	0.10	2.007	200.7	0.201	20.07
.34	4.28	37.0	1.456	12.59	.09	1.865	230	.168	20.72
.33	4.21	38.7	1.390	12.77	.08	1.717	268	.137	21.46
.32	4.14	40.5	1.326	12.95	.07	1.560	318	.109	22.29
.31	4.07	42.4	1.263	13.14	.06	1.396	388	.0837	23.26
0.30	4.00	44.4	1.200	13.33	0.05	1.220	488	0.0610	24.40
.29	3.925	46.7	1.138	13.53	.04	1.032	645	.0413	25.80
.28	3.85	49.1	1.078	13.74	.03	0.828	920	.0248	27.60
.27	3.77	51.7	1.018	13.96	.02	.603	1508	.0121	30.15
.26	3.69	54.6	0.959	14.19	.01	.345	3450	.0034	34.50
0.25	3.61	57.7	0.902	14.43	0.00	0	Infinite	0	Infinite

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TABLE 2

Values of the Correction Term G, in Formulas (13) and (14). As a Function of the ratio $n_1 \delta_1$ ($= n_2 \delta_2$) of the bare diameter of the wire to the winding pitch.

$n_1 \delta$	G	$n_2 \delta_1$	G	$n_1 \delta_1$	G	$n_1 \delta_1$	G
1.00	0.56	0.50	−0.14	0.30	—0.65	0.10	−1.75
.95	.51	.48	.18	.28	.72	.09	1.85
.90	.45	.46	.22	.26	.79	.08	1.97
.85	.39	.44	.26	.24	.87	.07	2.10
.80	.33	.42	.31	.22	−0.96	.06	2.26
0.75	0.27	0.40	−0.36	0.20	−1.05	0.05	−2.44
.70	.20	.38	.41	.18	1.16	.04	2.66
.65	.13	.36	.46	.16	1.28	.03	2.95
.60	.05	.34	.52	.14	1.41	.02	3.36
.55	−0.04	.32	.58	.12	1.56	.01	−4.05
0.50	−0.14	0.30	−0.65	0.10	−1.75		

TABLE 3

Values of the Correction Term H in Formulas (13) and (14), as a Function of the Total Number of Turns, n of the Coil.

n	H	n	H
1	0.00	40	0.315
2	.11	45	.32
3	.17	50	.32
4	.20	60	.32
5	.22	70	.32
6	0.23	80	0.33
7	.24	90	.33
8	.25	100	.33
9	.26	150	.33
10	.27	200	.33
15	0.29	300	0.33
20	.30	400	.335
25	.30	500	.34
30	.31	700	.34
35	.31	1000	.34

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TABLE 4

Values of the Equivalent Diameter Ratio for Calculations with Polygonal Coils.

Triangle		Square		Hexagon		Octagon		Twelve-sided	
$\frac{b}{D}$	$\frac{d}{D}$	$\frac{b}{D}$	$\frac{d}{D}$	$\frac{b}{D}$	$\frac{d}{D}$	$\frac{b}{D}$	$\frac{d}{D}$	$\frac{b}{D}$	$\frac{d}{D}$
0	0.827	0	0.900	0	0.955	0	0.974	0	0.989
0.01	.729	0.01	.844	0.01	.928	0.01	.958	0.01	.981
.02	.718	.02	.838	.02	.925	.02	.957	.02	.980
.03	.711	.03	.834	.03	.923	.03	.956	.03	.980
.04	.705	.04	.831	.04	.922	.04	.955	.04	.980
0.05	0.700	0.05	0.828	0.05	0.921	0.05	0.954	0.05	0.979
.06	.696	.06	.826	.06	.920	.10	.952	.10	.978
.07	.693	.07	.824	.07	.919	.15	.951	.15	.978
.08	.690	.08	.822	.08	.918	.20	.951	.20	.978
.09	.687	.09	.821	.09	.918	.30	.950	.30	.978
0.10	0.685	0.10	0.820	0.10	0.917	0.40	0.950	0.40	0.977
.15	.676	.15	.815	.15	.916	.50	.950	.50	.977
.20	.670	.20	.812	.20	.914	1.00	.949	1.00	.977
.30	.662	.30	.808	.30	.912				
.40	.657	.40	.805	.40	.912	$\frac{D}{b}$		$\frac{D}{b}$	
0.50	0.654	0.50	0.803	0.50	0.911	0.50	0.949	0.50	0.977
.60	.652	.60	.802	.60	.911	0	.949	0	.977
.70	.651	.70	.802	.70	.911				
.80	.650	.80	.801	.80	.910				
.90	.649	.90	.801	.90	.910				
1.00	0.648	1.00	0.800	1.00	0.910				
$\frac{D}{b}$		$\frac{D}{b}$		$\frac{D}{b}$					
0.50	0.646	0.50	0.799	0.50	0.910				
.00	.643	0	.798	0	.909				

SUMMARY: A set of tables for the simple calculation of the inductance of single layer helical and polygonal coils and their design is given; together with full illustrations of their use.

Union College,
Schenectady, N. Y.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

ISSUED DECEMBER 25, 1923—FEBRUARY 26, 1924

BY

JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, D. C.)

1,478,638—H. G. Cordes, filed May 10, 1917, issued December 25, 1923.

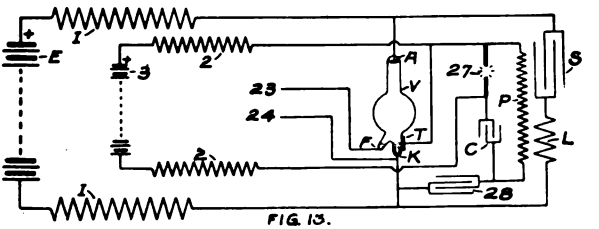

FIG. 13.

NUMBER 1,478,638—Electric Oscillstor

ELECTRIC OSCILLATOR of linear-sinusoidal characteristic in which oscillations are started by superposing a transient current upon a direct current and the oscillations sustained by means of a sustained alternating current. The oscillator comprises a direct current capacitance charging circuit, a direct current discharger circuit, a sinusoidal discharge circuit, and a uni-directional current relay discharger therefor. An auxiliary oscillating current relay discharger actuating circuit is coupled to the said discharge circuit for actuating the relay discharger of the oscillator.

1,478,806—A. F. Victor, filed March 1, 1922, issued December 25, 1923.

METHOD OF DELIVERING ILLUSTRATED LECTURES OR SONGS, comprising an arrangement at a radio broadcasting station of a manually operated picture projecting apparatus and a similar apparatus at a radio receiving station. The lecturer at the radio transmitting station produces a signal when he desires the next

*Received by the Editor, March 18, 1924.

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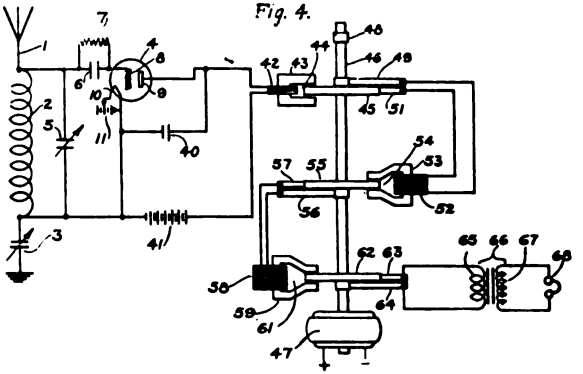
picture displayed, enabling the operator at the radio receiving station to make a similar change, so that the audience may follow the words of the lecturer by observing the displayed picture.

1,479,146—R. E. Marbury, filed March 5, 1921, issued January 1, 1924. Assigned to Westinghouse Electric and Manufacturing Company.



NUMBER 1,479,146—Radio Receiving System

RADIO RECEIVING SYSTEM wherein signals are amplified after rectification by means of a dynamo-electric machine, the incoming signals operating to produce a change in the magnetic flux of the machine for affecting the current output.

1,479,256—H. K. Sandell, filed August 28, 1919, issued January 1, 1924.

SPACE CURRENT DEVICE or electron tube wherein a block of insulating material is employed to support the electrodes with non-conductive spacing means supported by the electrodes independent of the block.

1,479,315—G. W. Pickard, filed January 12, 1921, issued January 1, 1924. Assigned to Wireless Specialty Apparatus Company.

ELECTRICAL CONDENSER AND PROCESS FOR MAKING THE SAME, in which the plates are formed by dielectric having thin film metal deposits on opposite sides thereof. The plates are stacked so that the metallic films contact with each other separated by the base dielectric material.

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1,479,475. O. Minton, filed January 19, 1923, issued January 1, 1924.

RADIO RECEIVING APPARATUS, in which the tubing system comprises two inductance coils connected in series and mounted for relative movement to vary the inductance linkage between them with a connection from a point of the circuit between the coils to the ground. The radio receiving apparatus which may be a radio frequency amplifier, detector, and audio frequency amplifier is directly connected across the terminals of the two inductance coils which are shunted by a variable tuning condenser.

1,479,638—V. K. Zworykin, filed August 28, 1922, issued January 1, 1924.

MULTIPLE REGENERATIVE LOOP ANTENNA AND CIRCUIT, in which one of the loops in the system is connected with the grid circuit of a detector tube and two or more loops are directly connected in the plate circuit of the tube forming a compact regenerative system.

1,479,778—H. J. Van Der Bijl, filed September 30, 1918, issued January 1, 1924. Assigned to Western Electric Company.

VACUUM TUBE DEVICE showing a construction wherein the plate electrode is formed in two parallel planes, the material of one plane being supported by metallic members which extend out from the other plane which in turn is substantially supported within the tube.

1,479,991—R. W. King, filed August 13, 1919, issued January 8, 1924. Assigned to Western Electric Company.

ELECTRON DISCHARGE DEVICE, in which the anodes are supported in the tube by forked metallic members mounted on the press. The tube structure also includes a block of insulating material supported by the anode, which insulating material in turn includes supporting connections for the grid and cathode.

1,480,208—W. G. Housekeeper, filed April 15, 1921, issued January 8, 1924. Assigned to Western Electric Company.

VACUUM TUBE, in which the filament electrode is supported by a spring suspension device anchored in an insulating member supported from the glass press of the tube.

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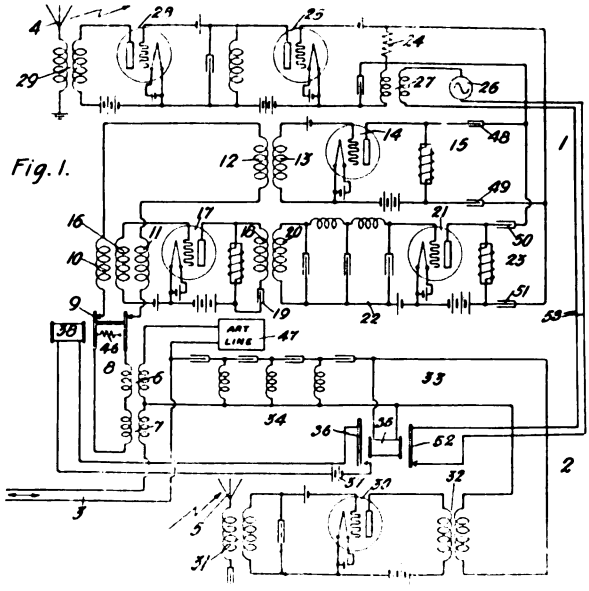
1,480,216—J. Mills, filed October 11, 1916, issued January 8, 1924. Assigned to Western Electric Company.



Number 1,480,216—Transmission System

Transmission System, in which a plurality of terminal stations comprising a radio telephone transmitting and receiving system and a line wire telephone system are employed. The object of the invention is to provide a circuit arrangement and system to prevent a circulation of power between the local transmitting and receiving system, or to insure that the signals received by a radio receiving system will be transmitted along the line wire in such manner as to not thereby effect the transmission of the same signal from the local radio transmitting station.

1,480,217 J. Mills, filed December 29, 1916, issued January 8, 1924. Assigned to Western Electric Company.

Method and Means for Signaling, wherein the intelligible interception of signals transmitted is prevented. This is accomplished by superposing on the outgoing talk or signal currents a complex noise formed by currents of several frequencies lying in the most important part of the voice frequency range. To clarify the signal at the proper receiving station, the invention provides for the superposition upon the incoming talk and its

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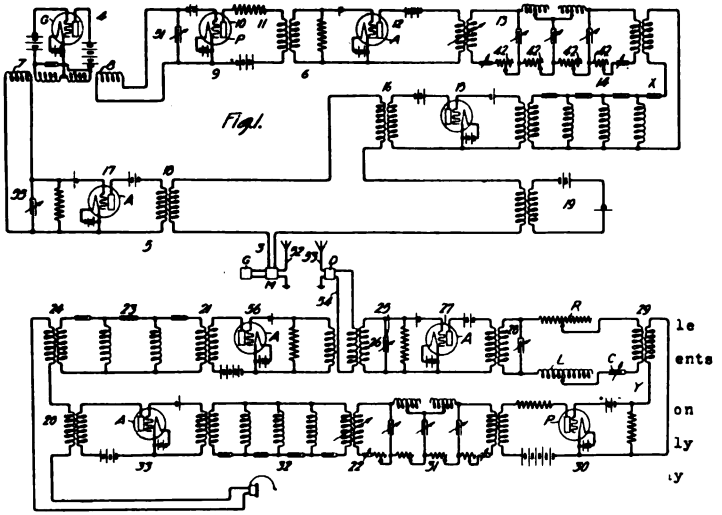
combined noise of a complex noise similar to the noise originally superimposed but exactly opposite in the phase of all its components. The noise is thus suppressed from the signal which may be received free from confusion.



NUMBER 1,480,217—Method and Means for Signaling

1,480,219—A. McL. Nicolson, filed June 25, 1917, issued January 8, 1924. Assigned to Western Electric Company.

VACUUM TUBE, in which the grid electrode is supported by a spring tension member which compensates for the expansion of the grid under changes in temperature in the tube whereby a uniform distance is maintained between the grid and the other electrodes at all times.

1,480,227—G. H. Stevenson, filed May 16, 1921, issued January 8, 1924. Assigned to Western Electric Company.

IMPEDANCE ELEMENT for use in high frequency alternating current circuits. The construction of the element is such that the value of its impedance may be accurately adjusted. An electrostatic shield is provided for the winding in the element.

1,480,388—A. L. Golden, filed September 16, 1919, issued January 8, 1924. Assigned to National Radio Company, of San Francisco, California.

OSCILLATOR for arc circuits in which the arc is formed between a pair of tungsten metal electrodes submerged in distilled water.

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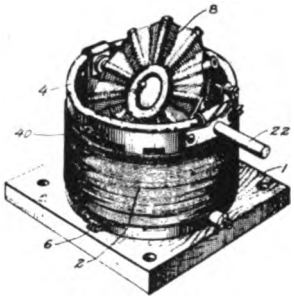
1,480,534—Alfred Crossley, filed August 16, 1922, issued January 8, 1924. Assigned to Hartman Electric Mfg. Company.



NUMBER 1,480,534—Variocoupler

VARIOCOUPLER having extremely small electroststic coupling between the windings. The variocoupler has a cylindrical stator and a flat rotor on which the secondary winding is arranged in such manner that the conductors are at all times approximately in a plane passing thru a diametrical line through the cylindrical stator. Sufficient electromagnetic coupling may be introduced to neutralize the effects of capacity coupling to secure substantially zero coupling.

1,480,604—W. Dubilier, filed January 27, 1921, issued January 15, 1924. Assigned to Dubilier Condenser and Radio Corporation.

TERMINAL CONNECTION FOR CONDENSERS, in which the conducting plates have terminals extending from the sides of the stack which are gripped together by a smooth round metallic hook.

1,480,659—Q. A. Brackett, filed December 7, 1920, issued January 15, 1924. Assigned to Westinghouse Electric and Manufacturing Company.

SIGNALING SYSTEM employing an arc which is controlled to produce signals with radiation of only a single wave. The arc is maintained in uninterrupted operation while the energy is supplied either to an oscillatory circuit or to a non-oscillatory circuit in the process of producing signals.

1,480,891—H. I. Becker, filed July 20, 1921, issued January 15, 1924. Assigned to General Electric Company.

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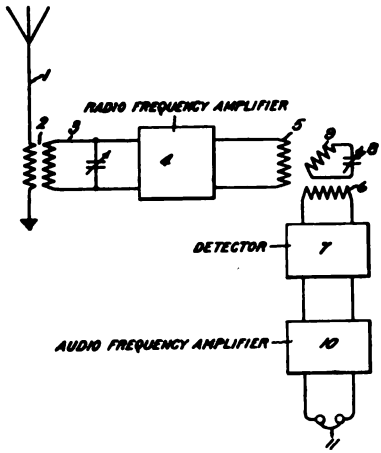

NUMBER 1,480,891—High Frequency Receiving System

HIGH FREQUENCY RECEIVING SYSTEM, in which a link circuit is provided between a radio frequency amplifier and the detecting circuit for preventing disturbing currents from being transferred from the amplification circuit to the detection circuit.

1,481,422—Gilles Holst, E. Oosterhuis, and J. Bruijnes, filed June 1, 1922, issued January 22, 1924. Assigned to Naamlooze Vennootschap Philips' Gloeilampen-Fabrieken, of Eindoven, Netherlands.

ELECTRIC DISCHARGE TUBE, in which the gas filling in the tube comprises neon to which 0.5 to 5 percent argon is added. The inventors state that the ionization potential of argon is lower than the potential at which the first inelastic collision in the first mentioned gas occurs.

1,481,945—J. Weinberger, filed March 3, 1921, issued January 29, 1924. Assigned to Radio Corporation of America.

RADIO RECEIVING SYSTEM, with a circuit arrangement between the antenna and the input of the receiver for eliminating interference. The system includes primary and secondary circuits with a pair of connections across the primary and secondary circuits, each connection comprising two parts including capacity and inductance in series, with one of the parts common to both connections and the other two parts arranged to couple the primary and secondary circuits. All of the factors in the circuits are adjustable to secure selectivity.

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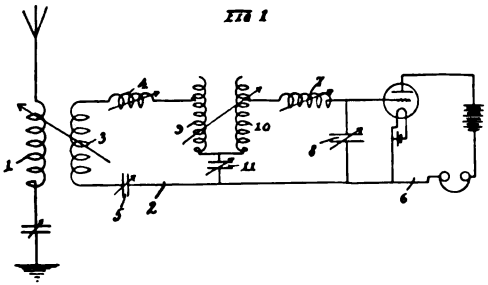

NUMBER 1,481,945—Radio Receiving System

1,482,122—W. N. Fanning, filed March 30, 1922, issued January 29, 1924.

RADIO SYSTEM for operating a signal or alarm upon the receipt of a certain definite call by radio. The device is particularly described in connection with a receiver on shipboard to sound an alarm for operation upon receipt of an "SOS" distress call.

1,483,383—H. K. Sandell, filed November 2, 1922, issued February 12, 1924. Assigned to Mills Novelty Company of Chicago.

RADIO RECEIVING SYSTEM comprising a loop antenna having two parts, one of which is connected in the grid circuit of a detector tube and the other in the plate circuit forming a regenerative system. The loops are spaced apart in parallel planes on the antenna frame and a condenser provided in one corner of the antenna frame which forms a variable tuning connecting means between the loops.

1,483,860—O. Von Bronk, filed September 3, 1921, issued February 12, 1924. Assigned to Gesellschaft für Drahtlose Telegraphie M.B.H., Hallesches, Berlin, Germany.

ANTENNA ARRANGEMENT FOR RADIO TELEGRAPHY, in which the antenna is supported by a conducting mast and an electromotive force applied to the antenna mast substantially equal and opposite to the electromotive force normally induced in the mast by radiation from the antenna.

1,484,189—J. A. Proctor, filed July 6, 1920, issued February 19, 1924. Assigned to Wireless Specialty Apparatus Company.

RADIO RECEIVING APPARATUS, utilizing a combined closed loop antenna and an open antenna. The open antenna comprises a conductor which is structurally separate from the turns of the

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loop antenna, but which is wound with the turns of the loop. The loop and antenna circuits are coupled with the receiving circuit.

1,484,269—E. Mayer, filed June 21, 1923, issued February 19, 1924.

METHOD AND ARRANGEMENT FOR PRODUCING NON-DAMPED OSCILLATIONS, in which a mechanical interrupter connected to a charging circuit is provided in circuit with a separate circuit arranged to receive energy from the charging circuit and produce oscillations. The charging of the circuit proceeds until the electrical energy stored in said circuit is such that when oscillations begin the momentary value or strength of the oscillation current is substantially equal and opposite with respect to the value of the charging current and the circuit then oscillates under such conditions that the charging current will be substantially equal to the amplitude of the oscillation current still existing after one complete oscillation. The circuit can then be broken without sparking and after a brief period of interruption and re-charging, more oscillations take place as before during a relatively longer period, and this operation can be repeated as long as oscillations are needed.

1,484,405—A. Oswald, filed January 12, 1920, issued February 19, 1924. Assigned to Western Electric Company.

SIGNALING SYSTEM having means for controlling a distant responsive device such as a call signal which will not respond to to false signals or electrical disturbances. A slow acting indicator is provided at the receiver which will only be actuated by the call signal. A source of opposing voltage is arranged at the receiver to cause the indicator to cease to respond upon cessation of the signaling waves.

1,484,411—H. S. Read, filed September 14, 1920, issued February 19, 1924. Assigned to Western Electric Company.

RADIO RECEIVING SYSTEM having a circuit arrangement for preventing radiation of the locally generated wave energy from the receiving antenna. The circuit arrangement includes a radio frequency amplifier, a local oscillator, and a detector and a circuit interconnecting the output of the amplifier and oscillator and the input of the detector which precludes radiation of energy from the local oscillator into the antenna.

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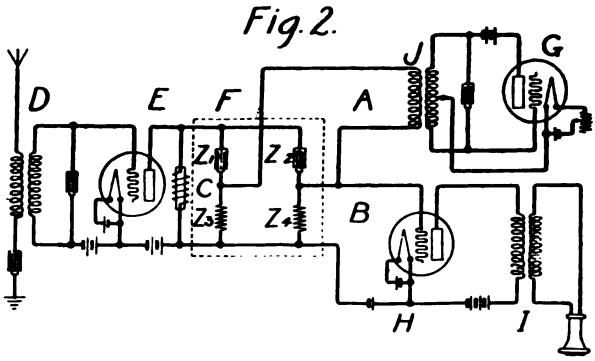
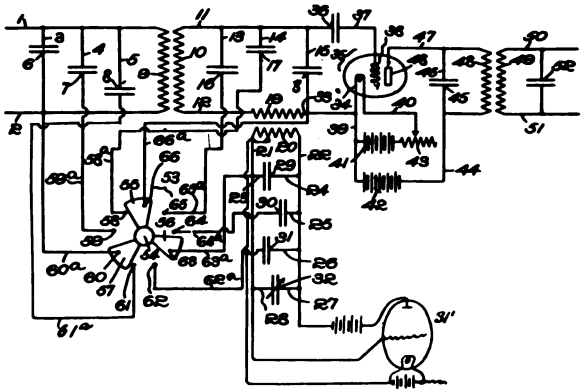


NUMBER 1,484,411—Radio Receiving System

1,484,605—J. H. Hammond, Jr., filed August 20, 1917, issued February 19, 1924.



NUMBER 1,484,605—Radio-Dynamic Receiving System

RADIO-DYNAMIC RECEIVING SYSTEM employing beat reception in which tuning is facilitated by an arrangement whereby two series of periodic impulses having different frequencies are impressed upon the oscillatory circuit. Periodic electrical beats are produced in the oscillatory circuit. The frequency of the beats may be maintained constant by varying the frequency of one of the series of impulses.

1,485,080—E. T. Jones, filed December 27, 1921, issued February 26, 1924.

SWITCH particularly designed for transferring connections from the antenna ground system to the radio transmitter or the radio receiver. The switch has a shaft mounted on the extremity

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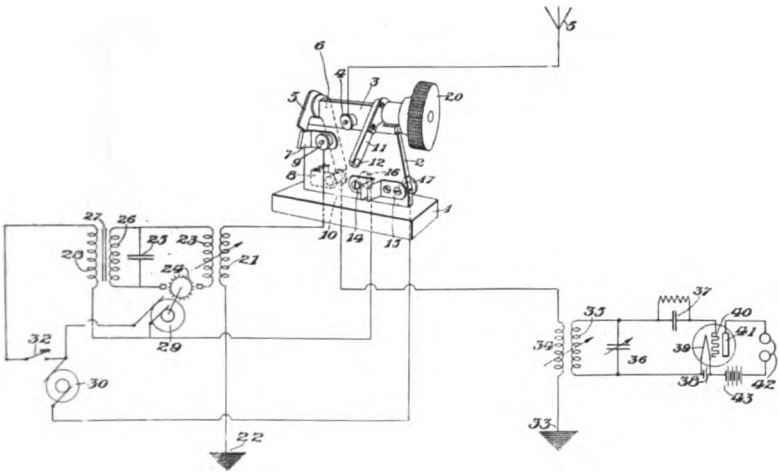
of an elongated support with switch blades carried by the shaft adapted to enter contacts on either side of the support when the switch shaft is rotated.



NUMBER 1,485,080—Switch

1,485,111—J. Bethenod, filed August 2, 1921, issued February 26, 1924.

RADIO TRANSMISSION SYSTEM, in which a plurality of radio frequency generators are provided at the same station for simultaneous operation to obtain multiplex transmission. The antenna is divided into two separate sections adapted to radiate different wave lengths. The generators are connected to these sections of the antenna and a connection of substantially infinite impedance provided in each of the sections to the wave lengths of the other section for limiting induction effects therebetween.

1,485,156—H. D. Arnold, filed August 28, 1917, issued February 26, 1924. Assigned to Western Electric Company.

SYSTEM OF DISTRIBUTION, in which an oscillation generator delivers current at a constant frequency, regardless of the nature of the load by reason of the fact that the load is prevented from reacting on the generator. An asymmetrically conducting repeater is placed between the output circuit of the oscillator and the load circuit permitting the transferring of alternating current to the load circuit, but preventing reaction from the load circuit upon the generator.

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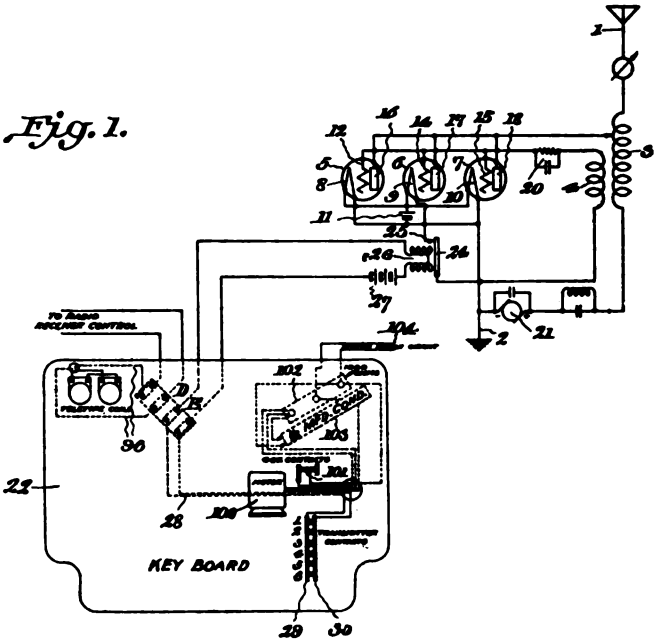
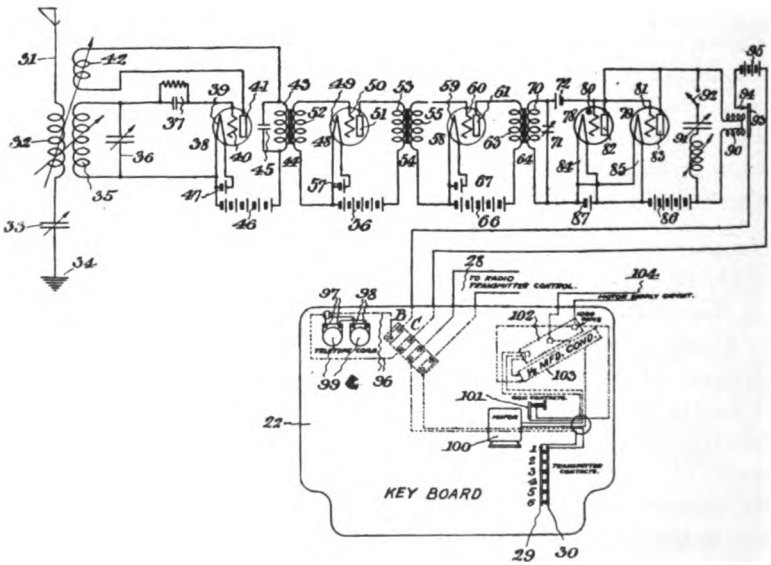
1,485,212—J. B. Brady, filed December 28, 1921, issued February 26, 1924. Assigned to Morkrum Company of Chicago, Illinois.

Fig. 1.





NUMBER 1,485,212—Radio Telegraph System

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Radio Telegraph System wherein signals are automatically transmitted and received directly in print. The system contemplates the broadcasting of news by radio from a central transmitting station to newspaper offices where the news may be received on a machine which automatically prints the signals. The system is also applicable in commercial radio for eliminating the human equation in handling heavy traffic. The invention introduces automatics in a practical way into radio.

1,494,087—F. M. Ryan, filed August 14, 1922, issued February 19, 1924. Assigned to Western Electric Company.

Radio Broadcasting Equipment for controlling the radio transmitter. The patent describes circuit arrangements between a studio and a radio transmission room whereby a person in either room may broadcast. Signal indicators are provided in each room to indicate automatically that the circuits are in operating condition for broadcasting from that particular room.

List of Radio Trade Marks Published by Patent Offiec Prior to Registration

(The numbers given are serial numbers of pending applications)

149,747—"Vactuphone" for telephone for aiding hearing. Globe Phone Company, Reading, Massachusetts. Claims use since April 12, 1921. Published December 25, 1923.

155,450—"UV" for vacuum tubes. Radio Corporation of America, New York City. Claims use since October 19, 1920. Published January 1, 1924.

157,776—"UC" for electrical condensers. Radio Corporation of America, New York City. Claims use since about August, 1921. Published January 1, 1924.

166,471—"Q-R" for complete radio sets and parts thereof. Robinson Specialty Company, Keyport, New Jersey. Claims use since on or about January, 1922. Published January 1, 1924.

170,674—"Test-Rite" for condensers. Scholes Radio Manufacturing Corporation, New York City. Claims use since August 14, 1922. Published January 1, 1924.

187,534—"Microstat" for rheostats. Premier Electric Company, Chicago, Illinois. Claims use since June, 1923. Published January 1, 1924.

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187,856—"TELOS" for radio receiving sets. Danziger-Jones, Incorporated, New York City. Claims use since February 3, 1923. Published January 1, 1924.

164,007—"RACO" in ornamental design for antennas. Ross Antenna Company, Providence, Rhode Island. Claims use since April 1, 1922. Published January 8, 1924.

170,764—"GEM" for radio apparatus. Gem Radio Company, Brooklyn, New York. Claims use since September 27, 1922. Published January 8, 1924.

187,634—"B-METAL" for radio receiving apparatus. B-Metal Refining Company, Detroit, Michigan. Claims use since January 15, 1923. Published January 15, 1924.

181,105—"LISTENING IN—THE WORLD OVER" in ornamental design for radio headsets. N. Baldwin, Incorporated, Salt Lake City, Utah. Claims use since February 28, 1923. Published January 22, 1924.

173,684—"THE STENTOR—THE VOICE FROM THE SKIES" in ornamental design for loud speakers. Stentorphone Company, Los Angeles, California. Claims use since August 1, 1921. Published February 12, 1924.

175,822—"T T" for loud speakers. John S. Timmons, Philadelphia, Pennsylvania. Claims use since about July 31, 1922. Published February 12, 1924.

186,609—"PERFECTONE" in ornamental design for radio receiving apparatus. Perfectone Radio Corporation, New York City. Claims use since October 1, 1923. Published February 12, 1924.

188,636—"MONODYNE" for radio receiving apparatus. National Airphone Corporation, New York City. Claims use since about May 15, 1923. Published February 19, 1924.

188,637—"GOLD GRAIN" for detectors and receiving sets. National Airphone Corporation, New York City. Claims use since October 12, 1922. Published February 19, 1924.

189,237—"NU-TRON" in ornamental design for amplifiers and tubes. Fred W. Brown, Bethseda, Ohio. Claims use since November 22, 1923. Published February 19, 1924.

189,884—"DICTO GRAND" for loud speakers. Dictograph

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Products Corporation, New York City. Claims use since March 6, 1923. Published February 19, 1924.

180,464—"ACME" for radio apparatus. Acme Apparatus Company, Cambridge, Massachusetts. Claims use since about March 1, 1919. Published February 26, 1924.

180,480—"THE LITTLE GIANT" for radio apparatus. Metropolitan Radio Corporation, Newark, New Jersey. Claims use since on or about August 1, 1922. Published February 26, 1924.

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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 12 JUNE, 1924 Number 3

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RECEIVING MEASUREMENTS AND ATMOSPHERIC DISTURBANCES AT THE BUREAU OF STANDARDS, WASHINGTON, NOVEMBER AND DECEMBER, 1923*

By
L. W. Austin
(Chief Radio Physical Laboratory, Bureau of Standards)

(Communication from the International Union of Scientific Radio Telegraphy)

During December, Nauen showed the lowest average morning strength which has been observed since the present methods of measurement have been used in the laboratory, while the afternoon signals have been of nearly normal strength.

All European stations measured in Washington show weakness in the forenoon at times during November and December. This may be connected with the European sunset, which, in the latitude of Nauen, occurs pretty early in the forenoon, Washington time, during the short days of winter. The corresponding effect in summer occurs in the afternoon, Washington time, and helps to produce the afternoon fading. This effect is much less pronounced in the case of Lafayette than in that of Nauen. This difference does not seem to be entirely due to difference in wave frequency, as St. Assise (UFT) shows less weakening than Nauen tho of nearly the same frequency.

The tables show that Lafayette continues to be approximately twice as strong as last year. Measurements made in Europe also indicate an increase in strength in this station, altho there have been apparently no changes except in frequency (12.8 kc. to 15.9 kc.) to explain the difference.

The atmospheric disturbances were somewhat stronger on an average than in the corresponding months of the year before.

* Received by the Editor, February 7, 1924.

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FIELD INTENSITY OF NAUEN AND OF DISTURBANCES ($f = 23.4$ kc., $\lambda = 12{,}800$ m.) IN NOVEMBER, 1923, IN MICROVOLTS PER METER

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
1	30.0	25	25.7	40
2	15.0	30	19.0	55
3	36.0	25	13.0	85
5	17.0	25	27.0	30
6	**	35	28.0	30
7	**	30	20.5	25
8	47.0	60	51.5	145
9	39.0	40	20.5	100
10	34.0	25	17.0	60
12	15.0	450	21.2	240
13	17.0	35	15.0	140
14	21.2	30	*	120
15	34.0	80	25.7	105
16	**	90	15.0	130
17	30.0	30	21.2	150
19	17.0	50	15.0	130
20	**	50	42.7	150
21	13.0	65	25.7	140
22	25.7	20	30.0	100
23	**	30	13.0	65
24	**	10	12.2	15
26	**	15	20.0	40
27	17.0	10	13.4	50
28	13.0	40	**	35
30	14.0	10	6.0	30
Average	24.1	52.4	20.8	88.4

*Not heard.
**Not sending.
....Not taken.

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FIELD INTENSITY OF LAFAYETTE AND OF DISTURBANCES ($f = 15.9$ kc., $\lambda = 18{,}900$ m.) IN NOVEMBER, 1923, IN MICROVOLTS PER METER

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
1	**	30	150	50
2	140	40	102	60
5	....	....	144	35
6	156	40	168	45
7	**	30	120	35
8	220	100	126	155
10	174	30	120	80
12	120	300	174	250
13	136	30	162	150
14	174	40	126	125
15	156	80	168	100
16	90	100	108	150
17	144	30	156	100
19	150	60	180	150
20	180	40	168	180
21	156	80	120	100
22	**	25	138	125
23	120	35	102	80
24	102	15	150	20
26	**	20	114	45
27	120	15	160	60
28	**	45	90	45
30	**	30	78	35
Average	146	55.2	135.8	94.6

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FIELD INTENSITY OF NAUEN AND OF DISTURBANCES ($f = 23.4$ kc., $\lambda = 12{,}800$ m.) IN DECEMBER, 1923, IN MICROVOLTS PER METER

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
1	7.5	20	19.0	45
5	2.0	40	13.6	60
6	4.0	40	7.0	35
7	2.0	25	5.0	45
10	2.0	15	19.0	25
10	12.8	40	7.0	25
11	6.0	40	14.5	55
13	5.0	15	13.6	40
13	4.0	30	**	40
14	9.5	30	22.0	25
15	13.0	15	7.0	95
17	17.0	15	21.5	30
18	8.5	20	....	....
20	....	....	**	20
21	5.0	25	18.0	50
22	2.0	25	17.0	40
24	7.0	15	....	....
26	**	10	34.0	30
27	2.0	20	25.7	40
28	21.2	10	43.0	20
29	17.0	15	55.7	40
31	6.0	25	21.5	35
Average	7.7	23.4	20.2	39.7

(Signal given as 2 when heard but not measurable.)

*Not heard.

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FIELD INTENSITY OF LAFAYETTE AND OF DISTURBANCES ($f = 15.9$ kc., $\lambda = 18{,}900$ m.) IN DECEMBER, 1923, IN MICROVOLTS PER METER

Date	9 A. M.		3 P. M.	
	Signal	Disturbance	Signal	Disturbance
1	60	25	119	50
5	84	50	108	65
6	102	45	96	35
7	196	30	102	50
8	**	20	138	30
10	125	45	90	25
11	125	50	78	60
12	125	20	102	45
13	**	40	108	45
14	132	35	90	30
15	120	20	222	85
17	**	15	**	40
20	....	....	108	25
21	90	30	180	60
22	102	25	102	45
24	**	20	....	....
26	**	15	180	35
27	102	20	180	40
28	**	15	212	25
29	163	20	180	50
31	132	30	114	40
Average	118.4	28.5	132	44

*Not heard.
**Not sending.
....Not taken.

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RATIO OF AVERAGES

Station	Signal P. M. / A. M.	Disturbance P. M. / A. M.	Reception Factor A. M. Signal / Dist.	Reception Factor P. M. Signal / Dist.
	NOVEMBER			
Nauen	0.87	1.68	0.46	0.24
Lafayette	0.93	1.71	2.64	1.43
	DECEMBER			
Nauen	2.62	1.68	0.33	0.51
Lafayette	1.12	1.54	4.15	3.00

SUMMARY: The signal strengths of the Nauen and Lafayette Stations, and the corresponding strength of the atmospheric disturbances in Washington are given for November and December, 1923.

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ON PROPAGATION PHENOMENA AND DISTURBANCES OF RECEPTION IN RADIO TELEGRAPHY*

By

F. Kiebitz

(Communication from the National Telegraph Engineering Bureau, Berlin, Germany)

The propagation of electromagnetic waves takes place in accordance with the principles of Huyghens and Fresnel. That is, every point which the wave reaches is to be regarded as the point of origin of elementary wavelets, which combine according to the usual laws governing interference, with due consideration to phase and amplitude.

If the propagation phenomena in radio telegraphy are to be described, as they take place on the separating surface between the earth and the atmosphere, it is first necesssary to know how these two media behave electrically. Experiments indicate that the atmosphere behaves practically as an insulator in which, therefore, electric and magnetic forces can exist. The earth, on the other hand, acts as a conductor. This is known because radio frequency alternating currents can be conducted thru the earth, and also because the electric forces and their various components are perpendicular to dry soil. This last effect can be directly observed; and in addition there is an indirect but particularly striking evidence for the vertical position of the electric forces relative to the earth's surface which depends on the following fact: A pair of horizontal antennas stretched in the same line near the surface of the earth and carrying electrical oscillation do not show any lateral radiation effect, whereas the same pair of antennas in free space, and therefore not oscillating in the neighborhood of the ground, act as a Hertzian transmitter or doublet in an exactly opposite fashion to the preceding pair and neither radiate nor receive waves in their own line of direction. This striking effect can be explained by assuming that the electric force is sensibly vertical to the earth's surface, whereas

*Received by the Editor, November 3, 1923. Translated from the German by the Editor.

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horizontal components of the electric field at the earth's surface do not have any effect.

If we consider the propagation of waves only over distances for which the earth's curvature may be disregarded, we shall make no great error if we regard the earth's surface as a limiting plane between an insulator and a conductor. Radio frequency alternating currents flow outward radially from the bottom of the antenna of the transmitter (or, to describe the phenomena in another way, radio frequency alternately positive and negative charges flow outward in concentric circles over the earth's surface from the antenna). The radio frequency alternating magnetic and electric fields in the air are inseparably connected with these currents or moving charges.

It is a matter of indifference whether we speak of charges which flow over the earth's surface or of electric fields which move thru the air, or of magnetic fields which grow and decay in the atmosphere, or of currents which flow in the earth's surface. We can employ interchangeably any of these pictures for a description of the same natural phenomena.

In the same way we may describe the excitation of a receiving antenna in various equally valid fashions. For example, we may concentrate attention on the charges which have come to the lowest point of the antenna, or on the currents which flow out of the earth's surface at this point, or on the electric forces which, traveling vertically, strike the antenna, or, finally, on the horizontal magnetic field which cuts the antenna as it moves forward.

Huyghens' principle determines the circular form of the wave front. The *decay* of field intensity with distance must be in agreement with that for an Hertzian oscillator, and therefore, must be proportional to the first power of the distance. As long as we are not forced to consider powerful absorption phenomena in the immediate neighborhood of the transmitter, no other description is possible. A direct quantitative determination of the law of attenuation would be of fundamental significance; but, so far as I know, this has never been carried out. The law connecting amplitude and distance has been obtained only indirectly thru absolute measurements of the field intensity, which latter have even been used for the determination of the effective height of a transmitting antenna and have given values which are in agreement with the conclusions drawn from the geometrical dimensions.

The curvature of the earth must be considered in connec-

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tion with the propagation of waves over larger distances. If the distance is small in comparison with the circumference of the earth, we may regard the earth's surface as very approximately spherical. Then, if A is the distance of the receiver from the transmitter measured on the spherical surface, the energy in spreading out over the arc A has not traversed a circle of circumference $2\pi A$ in this plane, but only a circle which is smaller in the ratio $\frac{\sin \alpha}{\alpha}$ (where α is the angle between the transmitter and the receiver viewed from the center of the earth), and, therefore, the field is increased in the ratio $\sqrt{\frac{\alpha}{\sin \alpha}}$.

When still greater distances have to be considered, and particularly in telegraphing to the opposite hemisphere, this well-known formula is no longer justified because of the curvature of the earth for non-directional transmitters. Only when telegraphing from pole to pole may we regard it as approximately correct, because the Huyghens' elementary wavelets then combine into an enveloping wave front which moves parallel everywhere to their lines of propagation or meridians, and which wavelets continue along meridianal paths on the opposite hemisphere because all points reached simultaneously on the paths of the waves are equally far from the poles. On the other hand, the further the transmitter is separated from one pole, the greater will be the difference of phase between the elementary waves which reach the distant receiver; and for a point on the equator, the path to the antipodes along the equator is 34 kilometers (21 miles) longer than that over one pole along the meridian.

If we consider, as an ideal case, that the earth is a perfect conductor and the atmosphere a perfect insulator, the wave which reaches the antipodes is obtained by summation of all the Huyghens elementary wavelets which have nearly equal amplitudes but various path differences of between zero and 34 kilometers (21 miles). Short waves will therefore combine practically to annul each other, because all those in one phase will be nearly exactly as powerful as those in the opposite phase, and therefore short waves are not suitable for communication over long distances.

For this reason, communication over great distances (considering the curvature of the earth) requires waves of the order of magnitude of 10,000 meters in length if audible alternating currents are not to be used for transmission. This fact has become well known thru the experience of those operating high-

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power stations, altho the curvature of the earth has not hitherto been given as the explanation for the necessity for long waves. It is perhaps not superfluous to emphasize this point because the desire to operate on short wave lengths is connected with efficiency of radiation from the transmitting antenna.

The above considerations cannot give more than the general order of magnitude of the necessary wave lengths. However attractive it may appear to be to calculate the most useful wave length for communication over the earth's surface between two distant points, it must not be anticipated that the results of a precise calculation will be substantiated by experiments, since the initial assumption that the earth is a uniformly good conductor of electricity is not sufficiently true in reality. The surface of the earth consists of water and land, and the waves have different velocities of propagation over these various surfaces. We can draw this conclusion in part directly from measurements of the stationary waves on wires stretched over the surface of the earth and in part indirectly from the deviations from a true direction or straight line path which occasionally result in directional telegraphy. The outward travel of the waves is more or less retarded by the differences between the non-homogeneous surface of the earth and a perfect conductor, and also because of the differences of path which result from the geometrical shape of the earth which leads to the elementary wavelets arriving at the distant receiver at different times. It is useless, however, to attempt to calculate these time differences because of the irregular distribution of water and land.

It is even more difficult to calculate numerically the changes in amplitude of the waves in traveling over a partially conducting surface, but even if such calculations could be successfully carried out, the influence of the atmosphere on the propagation phenomena would again alter the phenomena. This influence has been experimentally found to change with the weather and even with the amount of sunlight.

The investigations which have been published relative to these effects are so numerous that it is impossible to discuss even the most important of them. Attention has been concentrated properly on Dr. L. W. Austin's formula for the calculation of signal strength at a distance in radio telegraphy. In this formula it is assumed that the absorption depends upon the wave length, and the numerical value of the absorption is empirically determined. Furthermore, the formula is limited to the distant effect during the day and over sea water. It is natural with such

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limitations that only roughly approximate formulas can be obtained for the radio signal at distant points, since a generally applicable formula cannot be set up as long as the much simpler task of calculating the state of the weather is not possible.

Unfortunately, it is difficult to obtain information experimentally on radio signals at considerable distances, since such signals are so weak that they can hardly be directly measured by normal physical apparatus capable of giving objective or direct readings. Instead, a number of subjective or indirect methods of observation have been developed which do not always give even the order of magnitude of the signal, much less a quantitatively accurate value. As a result of this, even to-day, after using radio communication for twenty-five years, we are still not able to state whether the day or night values of signal correspond to normal travel of the waves. Partly as a result of the uncertainty relative to these fundamental questions, there have been propounded a great number of theoretically possible explanations which, however, sometimes border on the fantastic.

We have taken pains at the National Telegraph Engineering Bureau to devise receiving arrangements which enable the determination of received signal strength to be objectively carried out, and we have set up as a desirable condition that the transmitting station must not be required to send out special long dashes, but that we shall be able to make objective measurements on the signals of every audible station transmitting normal text. The measuring apparatus consists of a heterodyne receiver, and the signals are sufficiently amplified so that comfortable readable deflections are obtained on a thread electrometer even when Honolulu, the most distant station, is transmitting. The apparatus can be calibrated, and is set up at two locations, one in Berlin and the other 100 kilometers (62 miles) further north in Strelitz. The details of this equipment will be published by Mr. G. Anders. During the year we have succeeded in increasing the reliability of this measuring apparatus and its ease of calibration to such an extent that simultaneous observations on the incoming field strength of American stations agree to within 30 percent.

With these receivers we have concentrated on the signals from the American stations at Rocky Point (Long Island, New York) and Marion, Massachusetts, as a regular proposition, for a considerable time. In carrying out our plans to determine objectively the amplitude of the incoming waves as well as to specify their diurnal and annual variations in more precise fash-

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ion than is possible by the subjective methods, our thanks are particularly due to Doctors A. N. Goldsmith and L. W. Austin, whom we have interested in this work. In particular, Dr. Goldsmith has most accommodatingly arranged that all our wishes regarding the procedure at the Rocky Point Station, as required by these researches, have been fully carried out.

These observations have been carried so far that, in the near future, a complete description by Mr. Bauemler will follow. It is already possible to take a definite position relative to certain questions which I have mentioned above.

The received field strengths fluctuate by day and night. The highest values are obtained at night. We have compared these values with those which can be calculated from the distance of the transmitting station, its antenna current, and its antenna height, provided the spherical shape of the earth is neglected and absorption is not considered. The comparison shows that even the night values are generally smaller than these theoretical values. Peak observations, which exceed the theoretical value, frequently occur, but they exceed the theoretical value by less than thirty percent, that is, by less than the possible error of the observations. It is possible to conclude from this that there is no urgent necessity for explaining the large signal values at night by assuming reflections in the upper layers of the atmosphere. It is much more plausible to conclude that the propagation phenomena at night from America to Europe take place with relatively small disturbance over a good conducting ground surface in a homogeneous atmosphere which acts as a perfect insulator.

As a result of the numerous subjective observations, it has been known for a long time that the dependence of transmission on time of the day and season of the year is particularly marked for short wave lengths. The following average values obtained during the past few months have been secured for the 16,400-meter wave of Rocky Point (Long Island, New York, United States). During winter nights, the waves arrive with practically the theoretical or calculated amplitude. During the day in winter time, the amplitude diminishes to one-quarter of the theoretical value. The signal strengths during the summer nights are only slightly below the theoretical value, while the daytime values are greater in the summer than in the winter, the ratio between the average day and night strengths during the summer months being only one-to-two.

Because of the subjective impressions, the differences are

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generally given as much larger. The reason for this is actually psychological. Specifically, the atmosphere disturbances make it very difficult to determine the true strength of signals subjectively. These disturbances are particularly strong in the summer, and increase the difficulty of reception to such an extent as to cause the observer to underestimate the intensity of the signals.

Only partial information is available as to the source of the disturbances of radio reception. A portion of these disturbances undoubtedly originate in atmospheric electrical effects, and many useful researches have been carried out to explain the mechanism of the production of the disturbances. It is well known that the normal electric field in the atmosphere, as well as the magnetic field of the earth, is continually varying. Such variations appear with widely different magnitudes, and, for example, cause disturbing currents in the wire lines used for telegraphy and telephony. It is not to be expected that these electrical disturbances will fail to produce some effect on the delicate radio receivers. Actually, thay cause severe disturbances in the receivers because the radio receiver must be so much more sensitive than the equivalent arrangements used for wire communication.

It would be natural to attempt to establish a connection between the weather conditions and these atmospheric electrical disturbances. With this in view, we have carried out observations during the last year at the research laboratory at Strelitz not only on disturbances of reception but also on various meteorological and atmospheric electrical factors. However, no relation has been established between the intensity of the disturbances, nor yet the signal strength, and any of the phenomena taking place on the earth's surface (see Wiedenhoff, "Jahrbuch der drahtlosen Telegraphie," volume 18, 1921, part 4). Altho the results have therefore been negative, they have had the beneficial result of stimulating the devlopment of more powerful methods of studying disturbances of reception.

It must also be recognized that many disturbances of reception which have hitherto been ascribed to atmospheric disturbances unquestionably originate in man-made electrical installations. Reception has been repeatedly carried out in cities when, because of war conditions or strikes, all electrical equipment was at rest. Under such circumstances there were very few disturbances of reception.

The types of noise produced by the disturbances in the telephones have also been made the basis of their investigation. There are, for example, very rough disturbances which cause a

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crackling sound and are well marked in ink recorder reception. They appear with different intensities in receiving sets connected to antennas which are directional along different lines, and are particularly well suited for objective observation because of their intensity.

We next recorded these disturbances in Berlin and Strelitz, utilizing for this purpose the time signals from Paris and Lyons and taking down the signals on a usable cable siphon recorder. The receivers were therefore tuned to the same wave length and the time signals enabling direct comparison of the records for detecting possible simultaneity of the disturbances. It was found that a large part of the disturbances occurred simultaneously in Berlin and Strelitz. Following this, the same investigations were carried out in Hamburg, and then in Mr. M. Dieckmann's station at Gräfeling near Munich where Messrs. H. G. Möller and Gleissner very kindly undertook to carry out the observations. These investigations showed that even with the receiving stations separated by 600 kilometers (375 miles), a large part of the disturbances coincided. At this point Dr. A. N. Goldsmith courteously participated in these investigations by arranging to have the time signals from Lyons similarly recorded at the Riverhead (Long Island, New York) receiving station of the Radio Corporation of America. The records of the disturbances obtained in New York and Berlin also showed many coincidences, so that it was established that there are disturbances which cover a large portion of the earth's surface. A more complete description of these observations has been published by Mr. M. Bauemler in three articles on "Simultaneous Occurrences of Atmospheric Disturbances" ("Jahrbuch der drahtlosen Telegraphie und Telephonie," volume 19, 1922, part 2; volume 20, 1922, part 6; and volume 23, 1923, part 1).

These researches have therefore established the important fact that there are disturbances of radio reception which cannot originate in the vicinity of the receiver. It is, therefore, necessary to seek their source in distant electrical discharges in the atmosphere or else in still unknown cosmic phenomena or in equally unknown occurrences in the interior of the earth. This last possibility might be somewhat preferred since Mr. Barkhausen has found that iron, placed in a variable magnetic field, is not magnetized or demagnetized smoothly and steadily, but in jumps and finite steps. The earth has a magnetic field of its own which is subject to very appreciable variations. The masses of iron in the crust of the earth must therefore be subject to simi-

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lar variations in their magnetization. It is true that we have no exact knowledge of the characteristics of the iron-containing portions of the earth. Yet is may be possible, thru investigations of the electromagnetic disturbances taking place on the surface of the earth and simultaneously over large areas, to discover in the future whether they are caused by the Barkhausen effect in the iron-containing portions of the earth.

Berlin, October, 1923.

SUMMARY: The propagation of electromagnetic waves over the earth's surface is studied, particularly for very long distance transmission, for which it is deduced that short waves are inherently unsuitable.

Methods of quantitatively determining the strength of received signals are shown to be necessary; and some of the results obtained by their use in connection with diurnal and seasonal variations of transmission are given.

The simultaneous appearance of disturbances of reception at widely separated points is investigated and found to exist, and it is concluded that the source of such widespread disturbances may be in iron-containing portions of the earth's crust subject to varying magnetic fields and with a stepwise altered magnetization according to the Barkhausen effect.

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THE CAPE COD MARINE SYSTEM OF THE RADIO CORPORATION OF AMERICA*

By

F. H. Kroger

(Radio Corporation of America, New York)

It is the purpose of this paper to give a general description of the system and equipment employed in the development of a marine coastal station, representing the commercial application of the more recent improvements known to the art of radio telegraphic communication.

Marine radio communication may be divided into the following classes of service (in the order of relative degree of importance of the divisions):

(a) Distress Signals; (b) Navigation Information; (c) Ship Docking and Diversion Orders; (d) Public Radiograms; (e) Press Items, and (f) Medical Service. In explanation of this classification, distress signals include only the SOS calls which are given by the ship in distress, and these affect the shore station only when it may be called upon to relay weak signals, or to call other ships to the aid of the distressed vessels. Navigation information includes time signals, storm warnings, compass bearings, radio beacon signals, and the like, which are generally not handled by commercial shore stations in this country, but are sent out by the government without charge. Ship docking reports are given within a short distance of the harbor for the purpose of arranging pier accommodations. Ship diversion orders are given by the ship owners to the ship for the purpose of altering the ship's destination while en route. In this way, immigrant ships are quickly diverted to less congested ports; cargo ships which leave without exact destination can be ordered into ports offering the best market; or they may be diverted to secure better handling or port facilities when some emergency suddenly alters conditions. This service must be prompt, and therefore, it is most efficient when there is little relaying to do. This requires that the shore station have a long daylight range.

* Presented before The Institute of Radio Engineers, New York, November 7, 1923. Received by the Editor, December 6, 1923.

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Public radiograms are sent by individuals (and average fourteen words per message). These messages can be allocated as approximately 75 percent relating to business and 25 percent for social purposes. However, this allocation is subject to wide variations depending on the time of year, the class of ship, and the season of travel. The shore station is called upon to handle this traffic with heavy peak loads and yet to give prompt delivery. It might at first appear that the ocean traveler is not concerned so much with time, but the advice of one who has handled thousands of complaints is that the individual is indeed much concerned with hours and even minutes, and that facilities must be provided for the handling of the large files of messages in short periods. Press items are broadcast at scheduled times. Consequently the matter of news selection has been given considerable attention, and the press material is compiled by a Metropolitan daily and edited in the light of experience in this special field, national and local propaganda being deleted. The shore station must be equipped to reach ships which are within a day of the continent of Europe, as the demand for American news is very pressing. A medical service which was established recently provides for a ship obtaining professional medical advice on application to the shore station; this advice is free and originates in the Navy Marine Hospital.

Therefore we have the following requirements for a shore station which shall be capable of handling all classes of commercial service:

(1) A continuous watch must be maintained on all wave lengths allocated for traffic. (2) Provision must be made for receiving calls at the same time that traffic is being accepted. (3) The number of channels on each wave length shall be sufficient to take peak loads of traffic without undue delay. (4) The daylight transmitting and receiving range shall be at least to mid-ocean. These requirements may be met by the erection of several stations separated by considerable distances and in fact, this was the solution previous to the war. This method of providing facilities for handling marine traffic has caused an unnecessary amount of interference among the stations themselves and also to other classes of communication. Each station controlled its own affairs and was not amenable to a traffic head, and the result has been that even stations under the same ownership are often found to be causing interference by competing for a ship's traffic. Economic considerations point to a radio central station as a practical solution to the problem of handling

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large peak traffic loads and also covering a sufficiently large zone to give the service desired. In addition, it is of course obvious that there will be required small stations in the large harbors which can take care of vessels which are about to dock. Realizing the advantages to be obtained by diminishing interference, and the resulting economic gain in having a centralized system, the Radio Corporation of America decided to proceed with the development of such a system. The Cape Cod, Massachusetts, property was chosen as being suitable, and in the beginning of 1921 the first work was started. The initial step was taken to provide a channel on 136 kilocycles (2,200 meters). It was felt that this must be operated on continuous waves (c. w.) in order to obtain the desired range and not to interfere with other classes of radio traffic. About this time, the one-kilowatt tube was just emerging from the research laboratory of the General Electric Company. In order to obtain data as early as possible on the use of high voltage tube sets, two of these one-kilowatt tubes were used in a circuit oscillating directly into the antenna, with an alternating voltage of 12,000 applied to each anode. The set was installed at the Marion, Massachusetts trans-Atlantic station and the receiving equipment at the trans-Atlantic receiving station at Chatham. The results were quite gratifying, and the engineering department was asked to supply specifications for a somewhat larger transmitter. The next step was to use three of the one-kilowatt tubes, with the additional refinement of providing a constant source of high tension direct current for the plate supply, and to have the tubes generate oscillations in a "tank circuit" coupled to the antenna. By so doing the harmonics which were causing interference when operating the two-kilowatt high voltage alternating current set were eliminated and the frequency of the transmitter was not controlled by the constants of the antenna. This new set was installed and turned over to the Traffic Department in December, 1922, about seven months after the first set was put into operation. The operators at Chatham reported that no one heard the three-kilowatt set and insisted that the shift engineer at Marion give them the two-kilowatt set. Upon investigating the situation, it was found that the two-kilowatt set was radiating a band of wave lengths instead of a pure wave, and that the operators aboard ship had become accustomed to picking up station WCC at Marion without special care in accurately setting their tuners. Therefore instructions were given to try out the new three-kilowatt set for a given length of time in spite of the protests of the Chatham operators. At the

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end of two weeks, the new set was being used exclusively and without protest. When the old two-kilowatt set was substituted some time later, while some minor changes were being made in the three-kilowatt set, there was an immediate complaint from the ships. This experience was of great value to the engineers and gave them courage to proceed along the lines taken later with regard to the higher frequency channels. The next step was to provide two channels on the lower frequency, which was done by erecting a second antenna, and using the two-kilowatt transmitter on 130 kilocycles (2,300 meters). Thus, in the course of a year, two transmitting and two receiving equipments were installed and in operation on the lower frequency band allotted for marine service. Six months later a request was made to consider the feasibility of enlarging the system to include the higher frequency band for marine service. Transmission measurements of signals radiated from Marion showed that it was not economical to place the transmitters at Marion, whereas signals sent out from Chatham were efficiently radiated. To operate a transmitter of several kilowatts input to the antenna at 500 kilocycles (600 meters) at the same place that the receiving is done on 136 kilocycles (2,200 meters) involved receiving signals giving a field intensity of less than one microvolt per meter without interference from a transmitter which was giving a field strength at the receiving antenna of the order of 10 volts per meter or an energy ratio of a hundred-trillion-to-one. Field work was immediately started at the Belmar laboratory to determine a method of producing modulated radiation which would be satisfactory for crystal reception and would not cause interference in the 136 kilocycle (2,200 meter) receivers. The fears expressed in general that interrupted continuous waves (i. c. w.) would be too sharp and that they were not efficient for crystal reception caused work to be done on producing a radiation which covered a band of frequencies rather than a single frequency. Such an experimental set was assembled and tests were made using various widths of frequency bands, in connection with a test-receiving station at a distance and with ships picked at random. From our receiving engineers' report it was apparent that with the usual ship receiver using a crystal detector the change of width of frequency band of the transmitter when receiving on stand-by position did not have an appreciable effect on the intensity and sharpness of tuning and when receiving on loose coupling it was only when the broadening was increased beyond 3 percent that the effect began to be noticed. With the

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noticed increase in broadness of tuning, also came a decrease in the intensity of the signal. The reports from the ships were not at all coherent until the broadening was increased to 10 percent, when the report was in all cases that the tuning was broad. With 3 percent broadening, the reports varied between "very sharp" to "same as spark station." From this data it was apparent that if the operators used loose coupling, the tuning would be little sharper when receiving i. c. w. than for a good spark station and there would be increased intensity with i. c. w. in that they would receive all the energy on one frequency. The experience with the two- and three-kilowatt sets at Marion on 136 kilocycles (2,200 meters) was considered sufficient evidence that the ship operators would not have difficulty with tuning as soon as they learned of the advantages resulting. Therefore we proceeded to develop a method of modulating the continuous waves which would radiate on one wave length and would give crystal detector reception efficiently. The first step was to find the law of reception efficiency as the ratio of time-on to time-off of energy flow was varied. This variation plotted with audibility as taken by the shunt box method as ordinates and the average power as abscissas showed that when the energy is flowing one-third of the time the audibility is greatest for a given maximum amplitude of energy flow. Since, however, we should consider the quality of note this gives when heterodyning, we find that at this point the note is not musical and not until the energy is on one-half of the time is the note sufficiently musical to make copying easy and to be of value in reading thru interference and static. It was considered advisable to sacrifice plain detector reception to the extent of reducing signal intensity about 20 percent in order to get musical quality of signal when heterodyning. The result is a musical note which is reported as not mellow, but having a quality helpful in reading thru interference. As a result of the experimental work, specifications were written and the equipment installed last summer.

The system which began with one channel on 136 kilocycles (2,200 meters) has now been enlarged to two transmitters on this frequency with a spare set; three corresponding receiving channels; one transmitter for 500 kilocycles (600 meters) with a spare set and two corresponding receiving channels. Equipment has also been provided for increasing to two transmitters on the higher frequency when required, and space provided for the installation of a third as a spare.

In brief the elements of the station are:

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CONTROL SYSTEM—The complete control of the system located at Chatham. Here the operators are able to start, st and key any one of the transmitters which are at Marion, and t transmitters in the power house at Chatham. The distance fr Chatham to Marion is 50 miles (80 km.). One wire is used for t control of each transmitter. The controls are duplicated at ea receiving channel, and signaling lamps indicate when the tran mitters are in service. The land lines terminate here which gi direct service to the Broad Street office in New York City, a distance of 200 miles (320 km).

TRANSMITTERS—The transmitters located at Marion consi of a three-kilowatt set and two five-kilowatt sets, while the tran mitters located at Chatham consist of two five-kilowatt set The three-kilowatt set is shown in Figure 1 and the genera assembly of the five-kilowatt set is shown in Figures 2 and 3.

ANTENNA SYSTEM—The antennas located at Marion are tw inverted "L" antennas drawn from the two outside end masts supporting the trans-Atlantic antenna. The down lead for each antenna is vertically under the lower end of the flat top terminating at the tuning coil in the field. The energy is fed to the tuning coil by transmission lines which are six hundred and three hundred feet (183 and 92 meters), respectively for the two antennas. The coupling at the antenna coils is conductive and at the transmitter end it is inductive. The antenna at Chatham consists of four "T" antennas supported from a 350-foot (107 meter) tower. The down leads are brought into a common point and then to the tuning coil at the base of the tower. The housing for the outdoor tuning coil is shown in Figure 4. Figure 5 shows the general arrangement of the antenna.

GROUND SYSTEMS—The ground system at Marion consists at present of an incomplete overhead distribution system which feeds thru balancing coils to points on underground bus wires which bond together the ground system of the trans-Atlantic set. This is not efficient for the marine wave lengths and will be supplemented as soon as the changes now being made in the trans-Atlantic ground are completed. The ground resistance for one antenna is fourteen ohms and for the other ten ohms. It is expected that this will be decreased to five ohms each. The ground system at Chatham consists of a mesh of buried wires which are fed by an overhead distribution system. On account of the large area available the resistance of this ground is of the order of two to three ohms.

The energy is fed to the tuning coil by a transmission line

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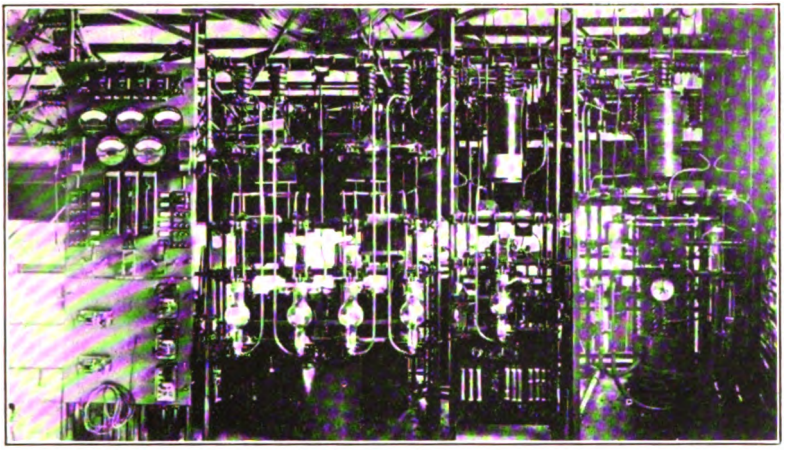

FIGURE 2

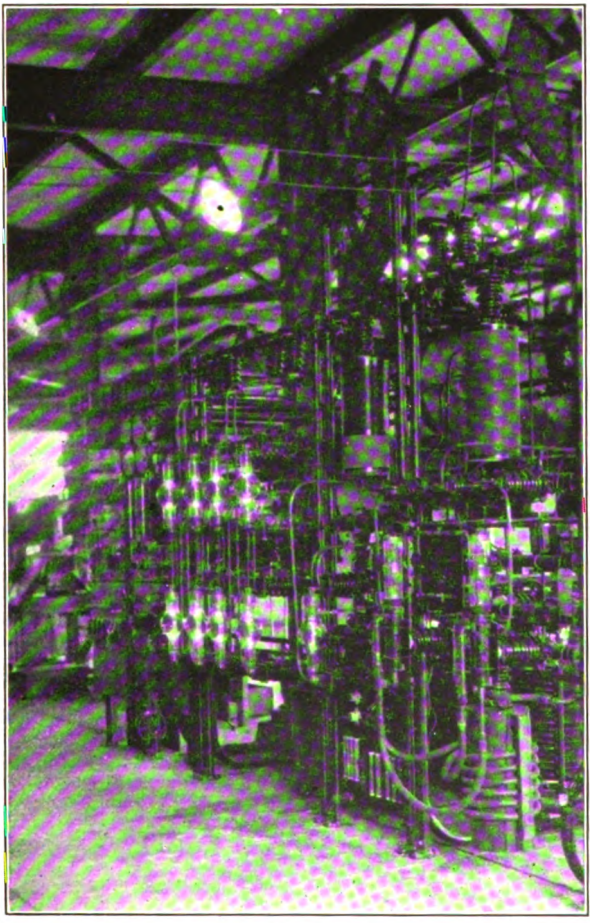

FIGURE 3

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900 feet (277 meters) long, and the coupling is inductive at both ends. The construction of this line is shown in Figures 6 and 7, and the latter figure shows the line leaving the power house where the transmittsrs are installed. The tuning house at the base of the antenna at Chatham is shown in Figure 8.

FIGURE 4

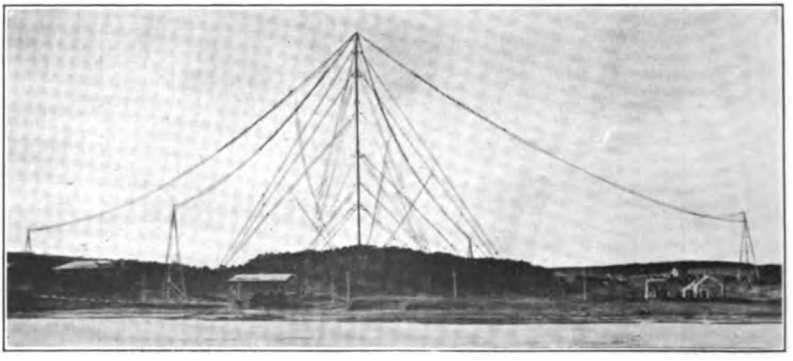

FIGURE 5

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RECEIVERS—There are provided three receivers for the lower frequency band and two receivers for the high frequency band. One of the complete receiving units is shown in Figure 9.

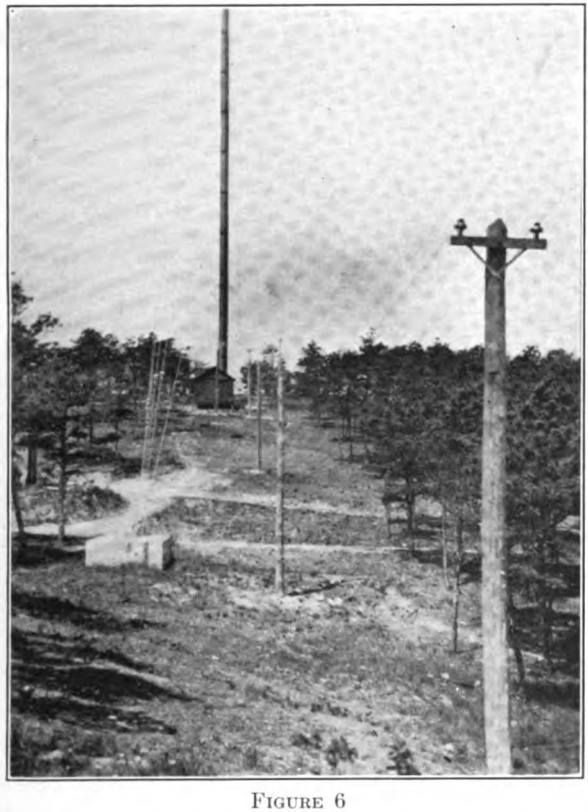

FIGURE 6

FIGURE 7

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FIGURE 8

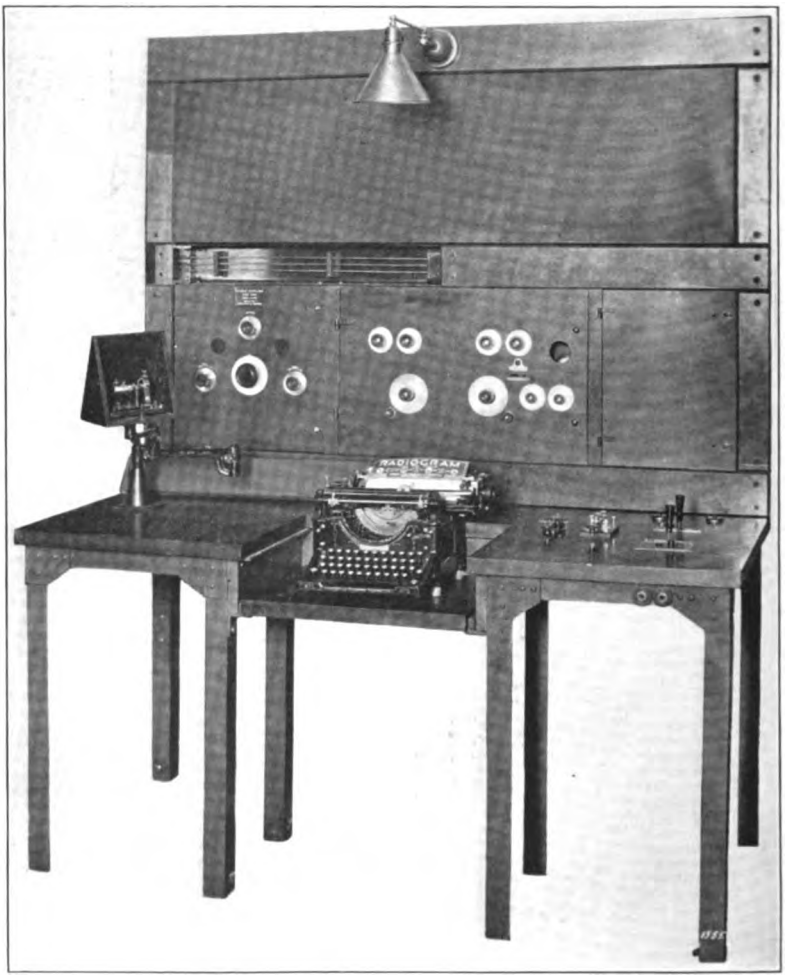

FIGURE 9

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RECEIVER ANTENNAS—There are two wave antennas, one downleads extending to the south, and the other to the west, which is used on the lower frequency band. The receivers are arranged so as to be used either all on one antenna or distributed between the antennas. The antennas for the higher frequency band consist of a vertical antenna and a loop system.

The results obtained from the Cape Cod system have demonstrated that the application of the recent developments in the art of radio to marine service is not untimely and will bring about an increased over-all efficiency in traffic handling and in the utilization of the limited available wave bands.

SUMMARY: This paper gives a general outline of the character of traffic handling which is required in a modern marine radio telegraph communication system. A description of the Cape Cod multiplex marine station of the Radio Corporation of America is given, and the operating problems encountered in supplementing spark transmitting sets with vacuum tube transmitters are outlined.

A brief description, with photographs of the transmitting and receiving equipment, is given; and the advantages made possible by the application of modern radio engineering methods to marine coastal stations are explained.

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KDKA

THE RADIO TELEPHONE BROADCASTING STATION OF THE WESTINGHOUSE ELECTRIC AND MANUFACTURING COMPANY AT EAST PITTSBURGH, PENNSYLVANIA*

By

D. G. Little

(Radio Engineering Department, Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania)

The purpose of this paper is to give a brief history of KDKA as the pioneer broadcasting station and to describe the equipment in use at the present time. KDKA was placed in operation on November 4, 1920, and broadcast as the first program the presidential election returns of that date. Semi-weekly programs were put on from that date until December 1st, 1920, when regular evening programs were commenced. The temporary call letters first assigned were 8ZZ. With a power output of only 100 watts, 8ZZ aroused great interest within a radius of 500 miles (800 km.) of Pittsburgh, and radio broadcasting as a public service may be said to have been started. On August 1, 1921, the power was increased to 500 watts and again on October 1, 1921 to 1,000 watts output.

A broadcasting station, in order best to serve the territory it normally covers, must have ample power; the quality both of reproduction and of programs must be of the best; and the service must be reliable. Provision should also be made for considerably increasing the power for emergencies or events of national interest. In order more nearly to approach these conditions, new and improved equipment has been added to KDKA both in the station and the studios. The station apparatus will be described in detail and the studio pick-up arrangement covered in a somewhat more general manner.

KDKA is now provided with three complete transmitting sets, designated as sets numbers one, two, and three.

*Received by the Editor, December 24, 1923. Presented before The Institute of Radio Engineers, New York, January 2, 1924.

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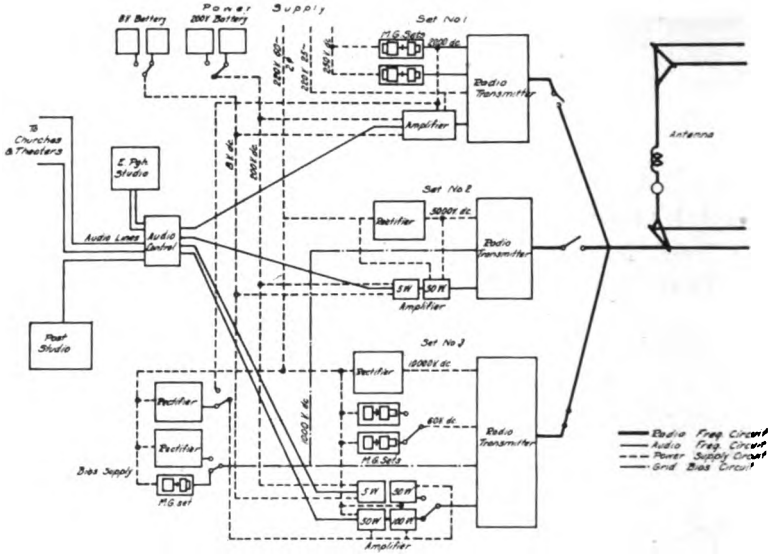

FIGURE 1—Schematic Diagram of K D K A, East Pittsburgh, Pennsylvania

A schematic diagram of the station is shown in Figure 1. It will be seen that each transmitter has its own speech amplifier and power supply apparatus. Set 1 has been fully described

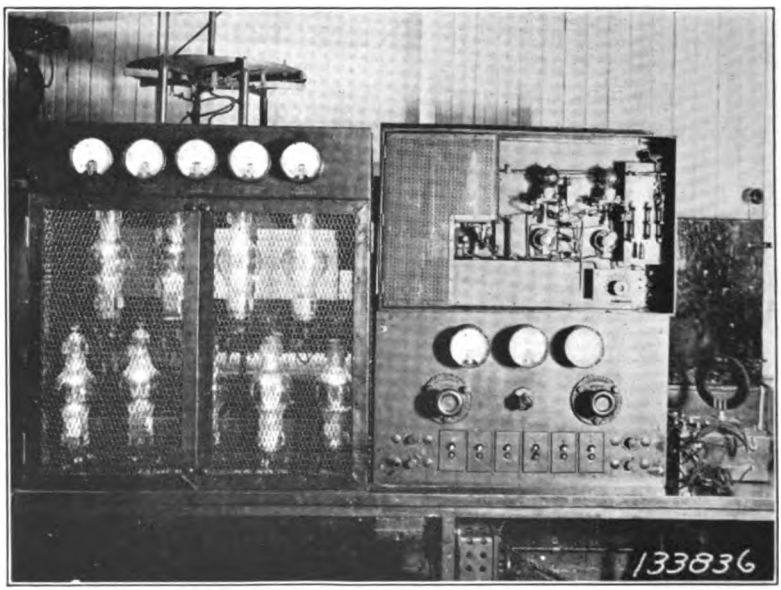

FIGURE 2—Number 1 Transmitter

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elsewhere, and the description will not be repeated here.[1] Figure 2, shows this set with its speech amplifier and power control panel. This transmitter was employed on regular programs up to October 1, 1923. Set 2 is shown in Figure 3. A single-phase

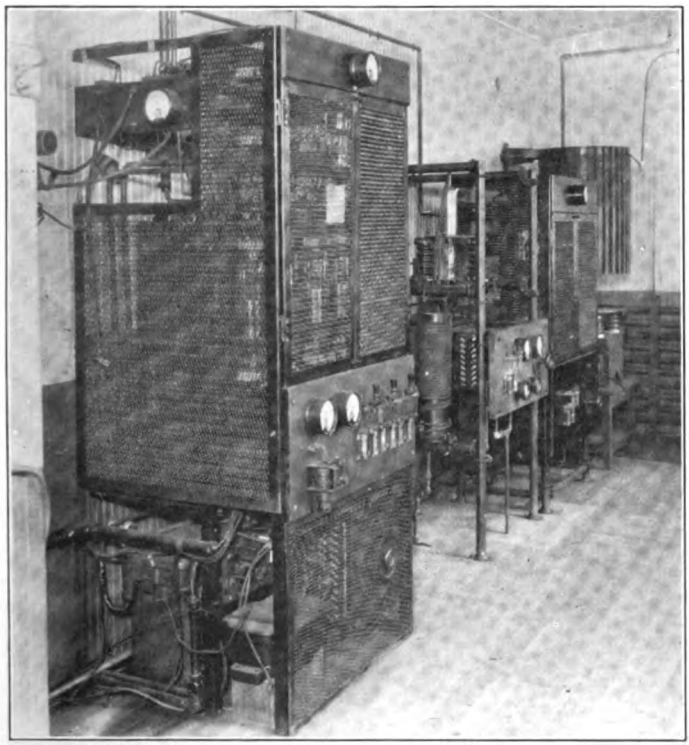

FIGURE 3—Number 2 Transmitter

60-cycle rectifier supplies the plate current at a potential of 5,000 volts to one metal water-cooled oscillator tube and a similar modulator tube. The constant current system of modulation is employed. The oscillator tube is connected to a primary circuit which in turn is loosely coupled to the antenna. This coupling system and also the speech amplifier is of the same design as that employed with the number 3 transmitter which will be described below.

NUMBER 3 TRANSMITTER

A general view of number 3 transmitter is given in Figure 4. The three panels at the left are high voltage rectifier and control

[1] "Electric Journal," June, 1922.

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panels while the radio set proper is mounted on the three panels at the right. It will be seen that this equipment is designed along standard switchboard lines, each unit being a separate panel. This construction gives accessibility together with a rugged and permanent design. Metal water-cooled vacuum tubes are employed as rectifiers, modulators, and oscillators in the usual manner, being supported on porcelain pillar insulators behind gray marble panels upon which are mounted the necessary instruments, water indicators, relays, and rheostats.

FIGURE 4—Number 3 Transmitter, General View

RECTIFIER

The rectifier for furnishing the high voltage direct plate current to the radio set is of the two-phase type. Primary power is 220 volts, 60 cycles, two phase, obtained from the works power plant. Four two-electrode tubes are used, two on each phase, as shown in the diagram Figure 5. The filaments of the rectifier tubes are connected in parallel and heated by 60-cycle power at 15 volts. The four rectifier tubes and associated apparatus have a maximum rated output direct potential of 10,000 volts The characteristics of a rectifier tube are given in Figure 6. In appearance the rectifier is the same as the three-electrode tube shown under Figure 14 below, except that the grid is omitted.

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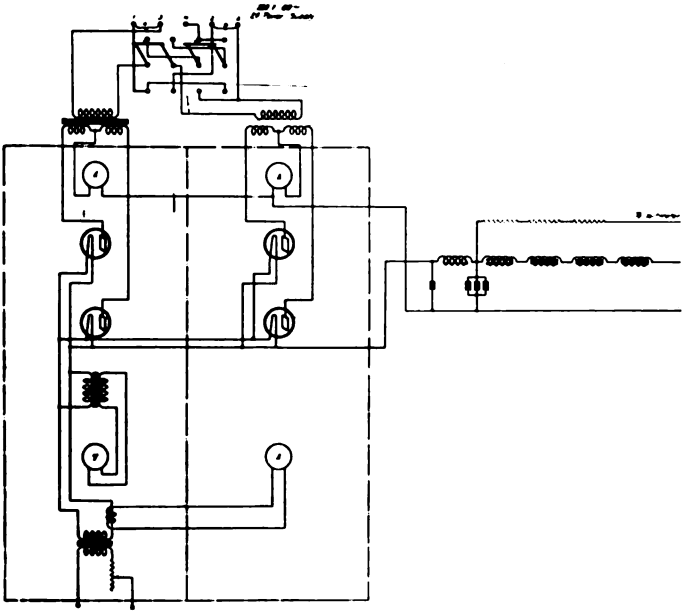

FIGURE 5—Two-phase Rectifier and Filter, circuit diagram

Each rectifier tube requires approximately 20 gallons of water per hour for cooling the plate, the water being fed into and away from the tubes thru ¼ inch (0.6 cm.) inside diameter rubber tubing twenty feet (6 m.) in length placed in the form of a coil, as can be seen in Figure 4. The tubing is connected directly to the works water mains and drainage system. The resistance of these two columns of ordinary city water is approximately two megohms, giving a leakage loss of 5 milliamperes at 10,000 volts or 50 watts. The total leakage loss from rectifier and radio set is thus not over 500 watts, which is less than the power required to operate a circulating and cooling system insulated from ground. This construction as described materially simplifies the equipment.

Figure 7 shows a close-up of the rectifier panels. The instruments on the panels are direct current ammeter for each phase, filament voltmeter, and ammeter reading total filament current. The filament meters are operated thru insulating transformers, as the tube filaments may be at 10,000 volts potential above ground. The primary power supply is stepped up by two transformers the secondaries of which are connected to the rectifier tube plates, as shown in Figure 5. The direct current ammeters for each phase are placed in the return leads to each transformer.

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Switches are provided so that the plate transformer primary voltage may be made 110, 154, or 220 volts by connecting to neutral and interphase voltage points for adjusting the plate voltage of the transmitter and thus its power output.

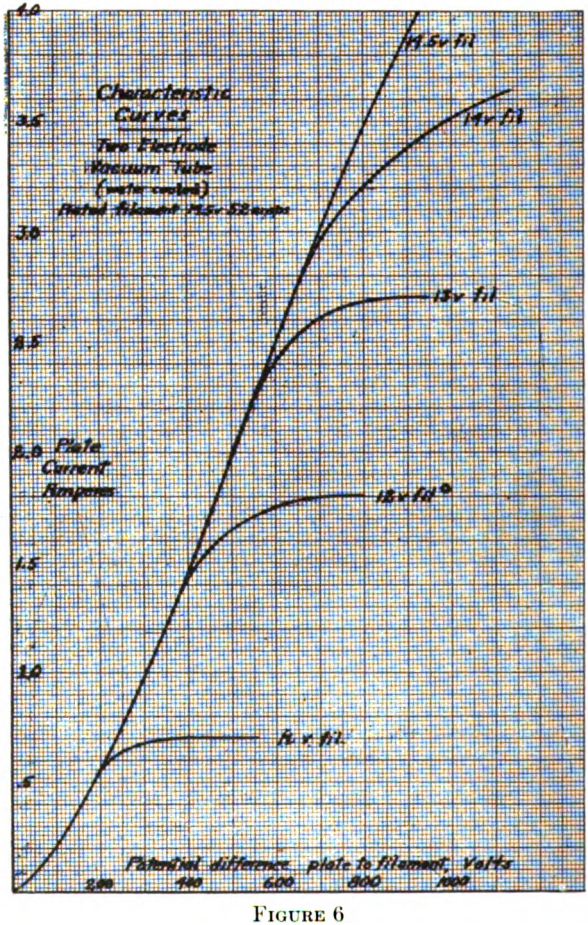

FIGURE 6

FILTER

A single step filter is employed for reducing the ripple voltage in the output of the rectifier. This consists of a 20-henry inductance connected in series with the positive lead, together with 4-microfarad and 12-microfarad condensers connected from either side of the inductance to the negative line. The constant current choke is composed of four 20-henry inductances in series. Figure 8 shows the filter as installed. This combina-

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tion of inductance and capacity leaves a ripple of less than 0.02 of 1 percent with the resonant frequency of the constant current choke coil and 12-microfarad condenser at 5.3 cycles, which is below the audio range.

The filter condensers deserve special mention in that it was necessary to design them for direct current operation. Four-

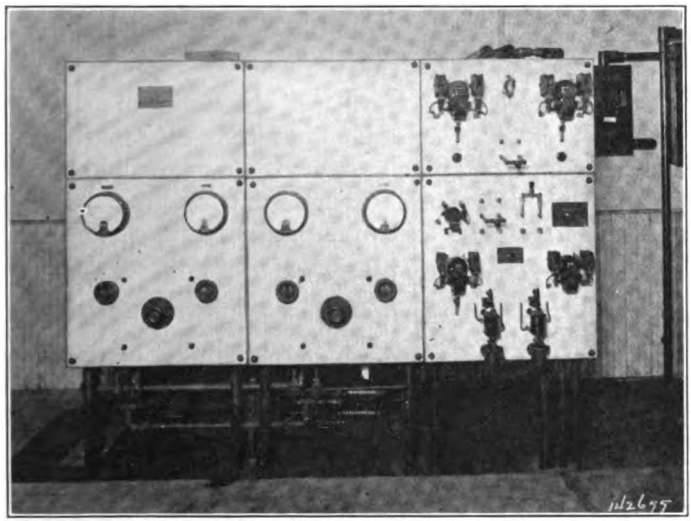

FIGURE 7—Two-phase Rectifier and Control Panels

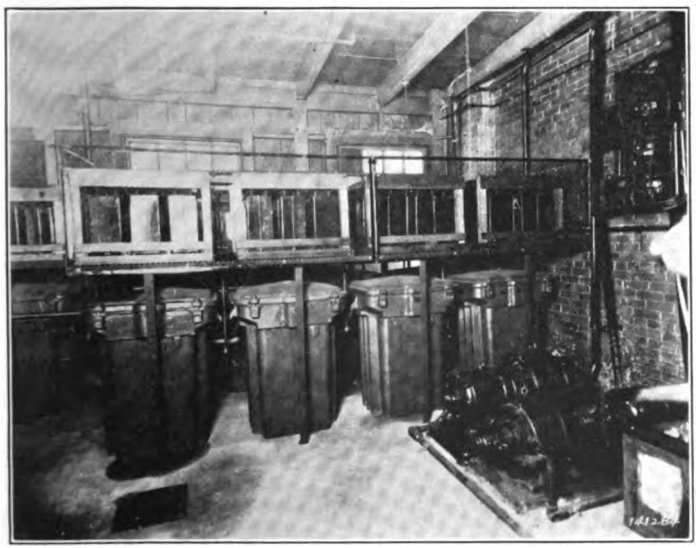

FIGURE 8—Filter Condensers and Choke Coils

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microfarad units are employed, each unit consisting of twelve separate 3-microfarad 3,500-volt condensers connected so that there are four sets of three condensers in series across the line. In order to divide the voltage equally across each of the three condensers, resistances of 300,000 ohms are connected across the line and tapped at 100,000- and 200,000-ohm points to the condenser series connections. From the filter the plate current goes to the radio set.

Radio Set

Figure 9 shows a close-up view of the radio set panels and Figure 10 the circuit diagram. Four three-electrode vacuum tubes are used, two as modulators and two as oscillators. The filaments of these tubes are connected in series and heated by direct current from either of two duplicate motor generator sets.

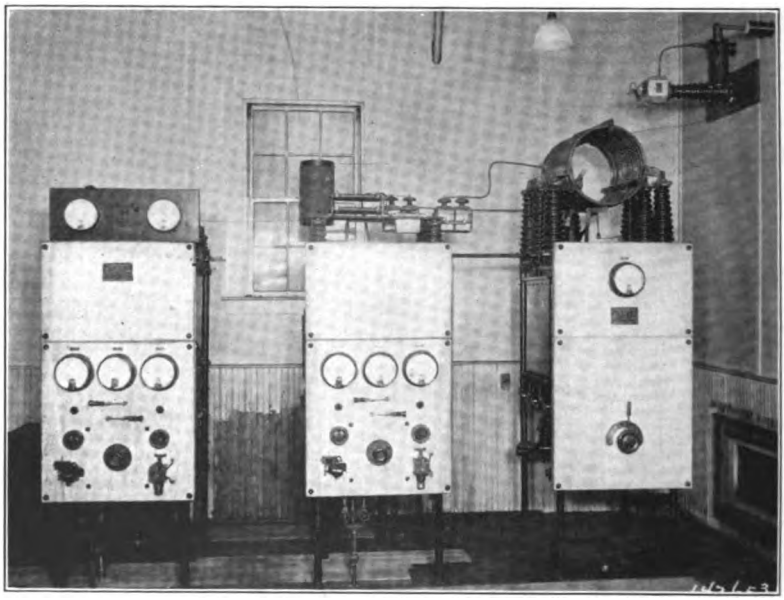

Figure 9—Number 3 Transmitter, Radio Panels

The oscillator tubes work into a primary radio frequency circuit, which in turn is loosely coupled inductively to the antenna.

Modulator Panel

The modulator panel is shown at the left in Figure 9. Because of the low impedance and relatively high static capacity of

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the tubes it is necessary to take steps to prevent the occurrence of local or parasitic oscillations. The plate current is fed to the tubes thru a radio frequency choke coil with input at the center of the winding and one tube connected to each end. Radio frequency choke coils are also connected with each grid to a single 20,000-ohm resistor connected in turn to the speech input transformer. This arrangement effectively prevents oscillations taking place between the tubes themselves or in series with the tubes, the plate choke coil, the grid input transformer, and ground. A

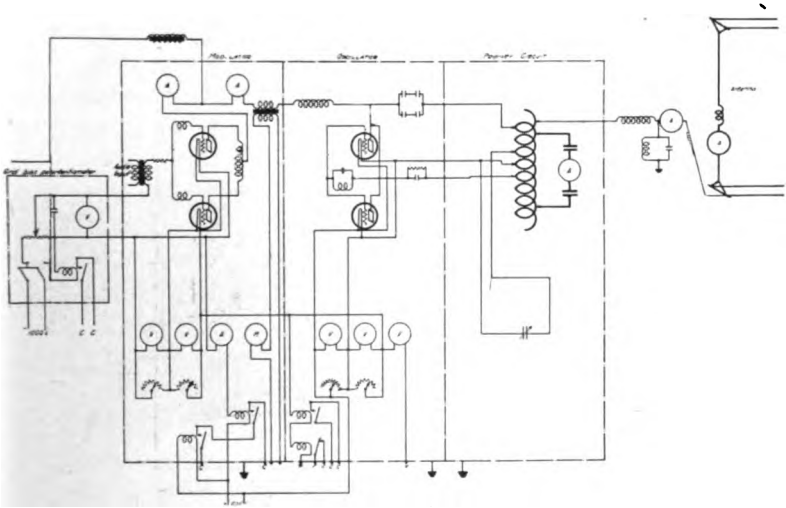

FIGURE 10—Number 13 Transmitter, K D K A, (Circuit diagram)

negative bias of 400 to 800 volts is required and is obtained from a small rectifier described below. Figure 11 shows the potentiometer employed for grid bias adjustment.

The modulator panel mounts filament voltmeters for each tube and rheostats for individual adjustment, modulation meter with its current transformer, and modulator and oscillator plate current meters which are of necessity connected in the positive direct current lead and hence mounted out of reach at the top of the panel. Water circulation indicators and control relays are also included. The modulator tubes each require 20 to 50 gallons of water per hour, depending upon the power used.

Cooling water for all tubes is obtained from a 100-gallon atmospheric pressure tank placed 20 feet (6 m.) above the tubes and supplied thru float valves from two separate water systems.

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Oscillator Panel

Figure 9 shows the oscillator panel in the center. The two tubes are mounted as before. Grid condenser and leak, plate coupling condenser, and radio frequency choke coil are mounted at the top of the panel as shown. The panel itself mounts fila-

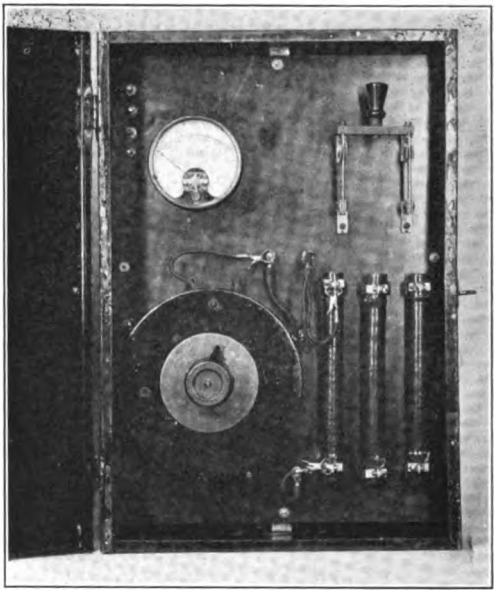

Figure 11—Potentiometer for Adjusting Modulator Grid Bias

ment voltmeters for each tube and plate voltmeter, together with water flow indicators and control relays. A choke circuit in series with the tube grids prevents parasitic oscillations. For cooling the oscillator tubes, 15 to 30 gallons of water each per hour is required.

The vacuum tubes employed as modulators and oscillators are of Westinghouse development and design. Figure 12 gives a photograph of a three-electrode tube. With a water circulation of 50 gallons per hour one tube will safely dissipate 6 kilowatts from its plate in addition to the filament energy of 750 watts. This tube is designed for a plate potential of 10,000 volts. At a plate efficiency of 66 percent 10 kilowatts output may be obtained. Figure 13 shows static characteristic curves. The alternating current plate impedance is 2,000 ohms, the amplification constant 13, and the saturation filament emission 8 amperes.

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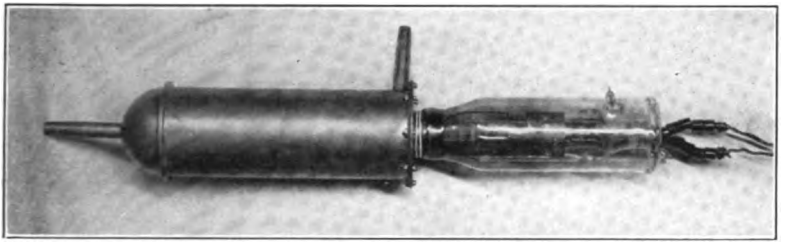

FIGURE 12—Three-Electrode Water-cooled Tube

PRIMARY CIRCUIT PANEL

The primary circuit panel is shown at the right in Figure 9. The circuit consists of two 0.001-microfarad oil dielectric condensers connected in series, and an inductance which may be seen above the panel. The condenser series connection provides a point of approximately ground potential for the circulating current ammeter shown mounted on the panel. A variable air condenser of 200 micro-microfarads maximum capacity is connected across two turns of the inductance and provides for vernier tuning of the primary circuit to 920 kilocycles (326 meters). Grid, plate, and ground taps are made from the inductance to the oscillator panel and coupling tap to the antenna.

ANTENNA

The radiating system is of the inverted L type, consisting of duplicate antenna and counterpoise shown in Figure 14. The antenna is made up of two round cages 170 feet (52 m.) long, 8 inches (20 cm.) diameter of eight number 14 wires* each, supported on spreaders 20 feet (6.1 m.) apart at a height of 90 feet (27.5 m.) above the station and 210 feet (64 m.) above the ground. The station end is supported by a steel mast 90 feet (27.5 m.) above the station roof, while the opposite end is attached to a brick smokestack. Separate ropes from each cage run over pulleys and down the sides of the stack with counterweights attached at the lower ends thus giving a uniform tension on the cages at all times. The counterpoise is of the same size and construction placed 110 feet (33.6 m.) below the antenna. A cage type down-lead is connected to the station end of the antenna and is made up of sixteen number 14 copper wires on 3-inch (7.6 cm.) diameter spacers. The down lead runs directly to a small inductance located on the roof of the station, the circuit continuing from a flexible tap on the inductance to the

*Diameter of number 14 wire = 0.064 in. = 0.163 cm.

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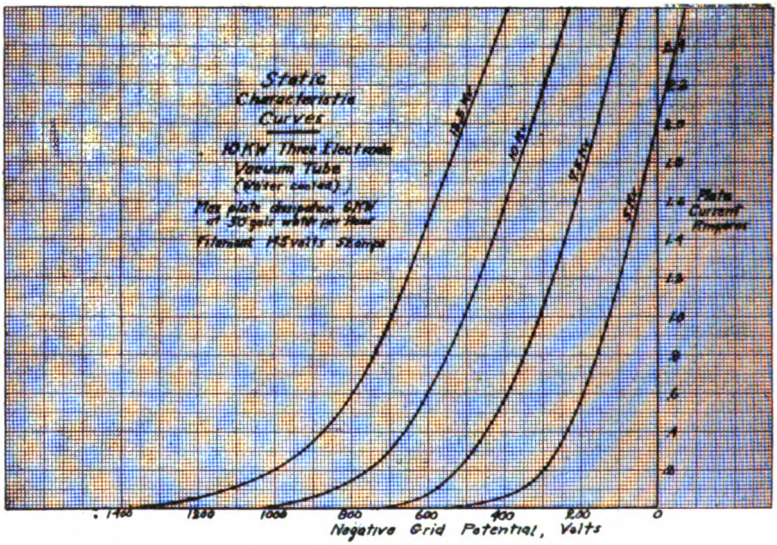

FIGURE 13

antenna ammeter, and thus thru a similar cage lead to the counterpoise. It will be noted that the circuit carrying the antenna current does not enter the station. Energy is supplied to the antenna thru a coupling lead attached to the station end of the counterpoise. The radiating system is tuned approximately to a frequency of 920 kilocycles (326 meters wave length).

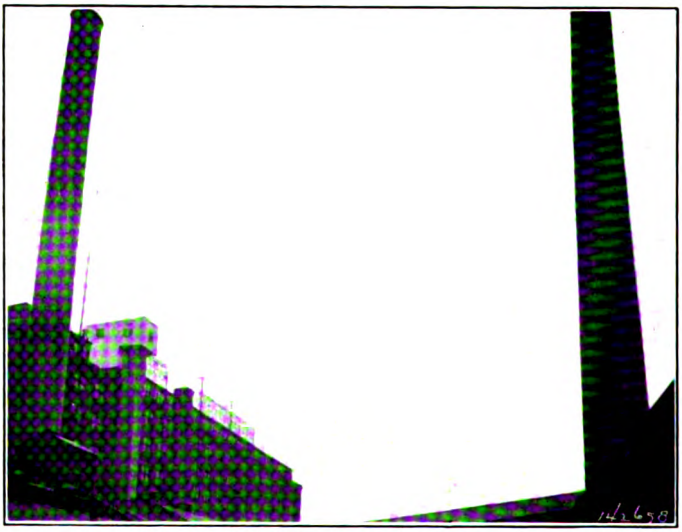

FIGURE 14—Antenna and Counterpoise

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The antenna coupling lead is connected to the primary circuits of any of the three transmitters thru inductance coils, the coil for number 3 set being shown at the upper right in Figure 9. The reason for employing an inductance in place of the usual coupling condensers for this purpose is to reduce the harmonics radiated from the coupling lead. The coupling condenser offers less impedance to a harmonic than to the fundamental frequency, whereas the inductance offers a greater impedance, the inductance giving four times the impedance of the condenser at double the fundamental frequency, nine times at triple frequency, and so on. In order still further to reduce the harmonic radiation, a parallel resonant circuit tuned to 920 kilocycles is connected between the coupling lead and ground, thus acting as a shunt to all frequencies except the fundamental of 920 kilocycles. The antenna system at KDKA has a total effective resistance of 16 ohms at 920 kilocycles.

Frequency Standard

With the large number of broadcasting stations operating at present, having assigned frequencies within ten kilocycles of each other, it is absolutely necessary that the radiated frequencies be correct to within 0.1 of 1 percent with practically no variation from night to night. A frequency standard for station use has been developed which allows this close regulation to be obtained. Figure 15 shows the external appearance and internal construction of the meter and Figure 16 the resonance curve for one of the meters used at KDKA, which is set for a frequency of 920 kilocycles. It will be seen that an adjustment of less than 1 kilocycle in 920 may be obtained. These frequency standards are designed to be adjusted to the frequency assigned to the station where they are to be used and sealed at the Bureau of Standards. All of KDKA's transmitters are provided with vernier tuners and frequency standards to enable precision adjustment.

Control

Number 3 transmitter is started and stopped by push buttons located on the attendant's desk, the operator's desk, or the control panel, thru contractors on the control panel shown in Figure 7. The attendant's desk may be seen in Figure 4. The operation of starting number 3 set for a program is as follows:

The station attendant makes an inspection of the set and then presses the water-control button, which by means of an electric-

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ally operated valve turns on the cooling water to all tubes. When the water pressure comes up to normal, a green signal lamp is lighted at all control stations. The operator may then start the set by pressing the power button. The plate current cannot

FIGURE 15A—Frequency Standard, interior

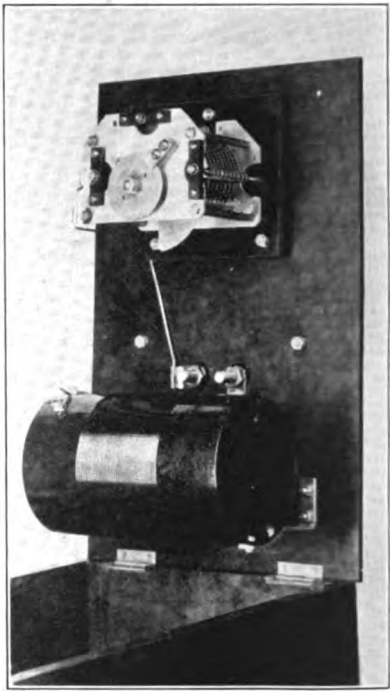

FIGURE 15B—Frequency Standard, exterior

be applied unless the water is on. Pressing of the power button at any control station starts the motor of the filament-heating motor generator set. When the filament generator has built up and the radio set filaments are lighted, a voltage relay closes a contact, which in turn closes contactors on the control panel, turning on the rectifier filaments and primary of the plate transformers. The radio set starts as soon as plate voltage is obtained from the rectifier and filter. When the primary of the plate transformers are connected to the line a red signal light appears at all control stations. The negative lead from the rectifier is connected thru an overload relay on the oscillator panel which opens the control circuit and shuts down the set in case the plate current exceeds a safe value. In case of filament

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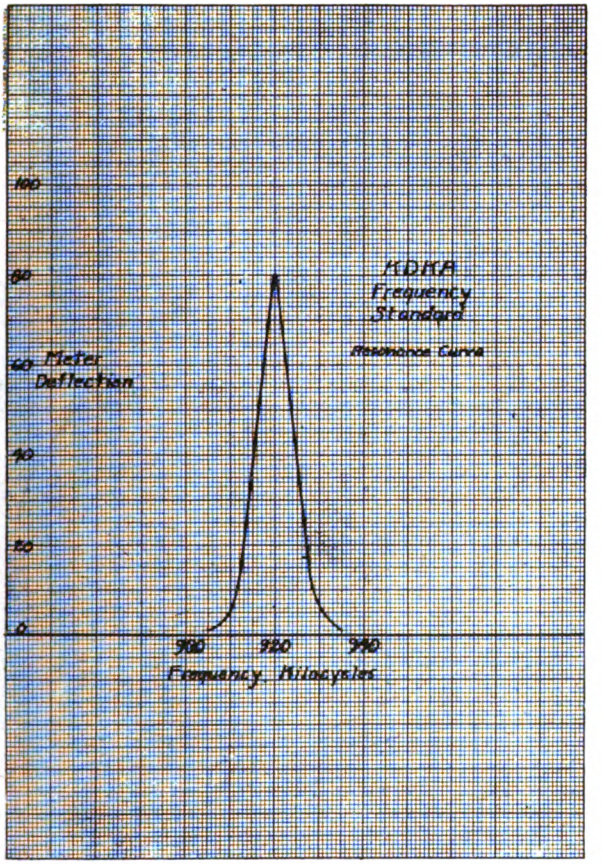

FIGURE 16

failure of one of the radio set tubes, all the filaments cease to burn because of the series connection. To prevent the plate voltage rising to abnormal value, a relay with coil in series with the filaments opens the control circuit and shuts down the rectifier and plate supply. In case the water supply fails, the set is also automatically shut down. The set is normally stopped by pressing the stop button for either water or power at any control station. This interlocked control system protects the vacuum tubes from injury and prevents mistakes on the part of the operators. Figure 17 shows a schematic diagram of the arrangement.

GRID BIAS AND AMPLIFIER PLATE VOLTAGE RECTIFIERS

Figure 18 shows two full-wave rectifiers and filters mounted in one frame. One supplies grid bias voltage for sets

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2 and 3, and the other power for the 50-watt speech amplifiers described below. Two 0.25-ampere glass rectifier tubes are used in each, the grid bias requiring approximately 1,000 volts direct current and the amplifiers 2,000 volts at the rectifiers. These rectifiers are started automatically by contactors from the control circuits of sets 2 and 3. As auxiliary supply, a small motor generator set is available for grid bias, while the amplifiers may obtain their plate current from the 2,000 volt generators connected with set 1.

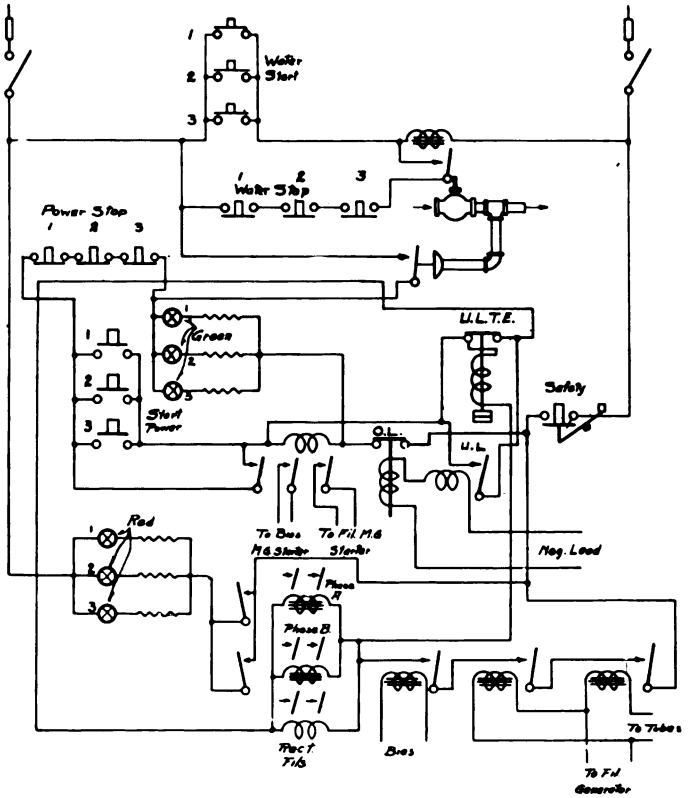

FIGURE 17—Number 3 Transmitter Control System

AUDIO SYSTEM

Two outside telephone lines, the "Pittsburgh Post" studio line and the East Pittsburgh studio line, terminate at the station operator's desk. The audio frequency circuits are shown by light solid lines in Figure 2. The station operator is in complete control of the station from this desk.

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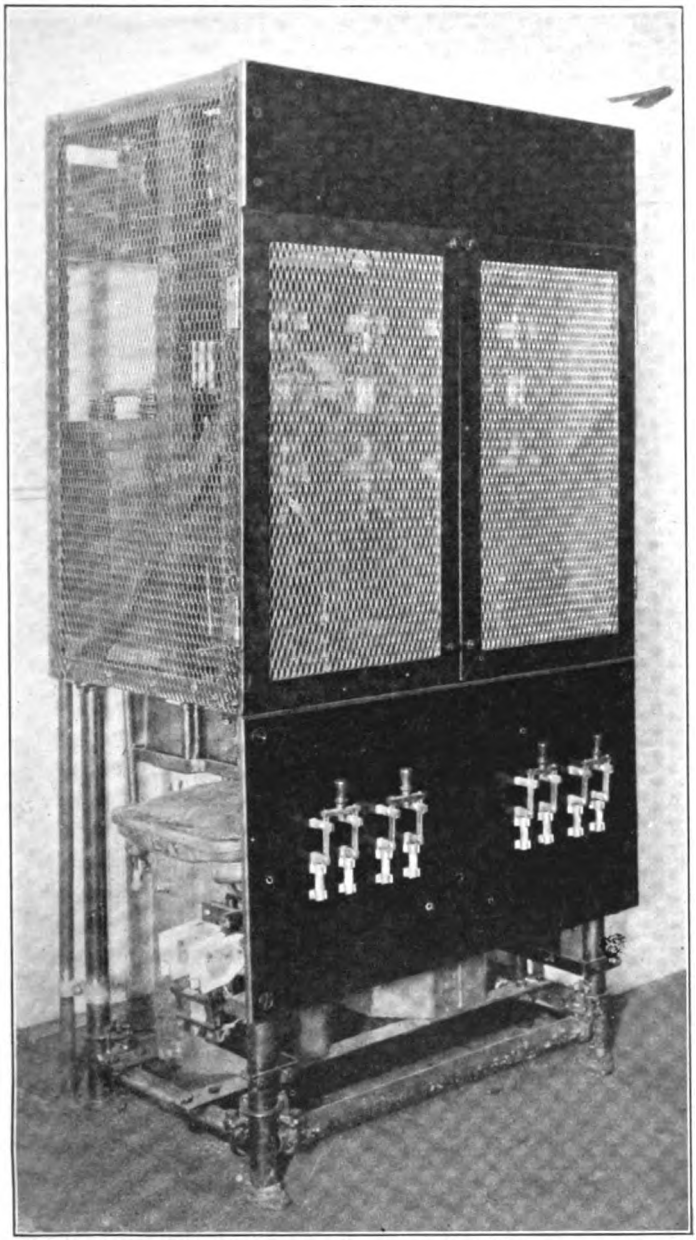

FIGURE 18—Rectifiers, Supplying Modulator Grid Bias and Amplifier Plate Voltages

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He may switch any line to any set, start and stop any set, and control the amount of modulation as indicated by the modulation meters at the desk. The operator may also, by means of switches and head receivers, listen in on any line, listen to the output of any amplifier, or by means of a crystal receiving set, listen to the radio output. Centralizing the control in this manner has been found to eliminate mistakes and generally improve the operation of the station.

The audio amplifier for set 2 consists of a 5-watt push-pull stage and a 50-watt push-pull stage, all transformer coupled. The number 3 set has a similar amplifier arrangement and also a 50-watt push-pull first step, and 100-watt push-pull second step which may be used in place of the 5- and 50-watt combination.

Figure 19 gives a photograph of the 50-100-watt amplifier. Storage batteries are provided in duplicate giving voltages of 200 for the 5-watt plates and 12 volts for the 5-watt and 50-watt filaments.

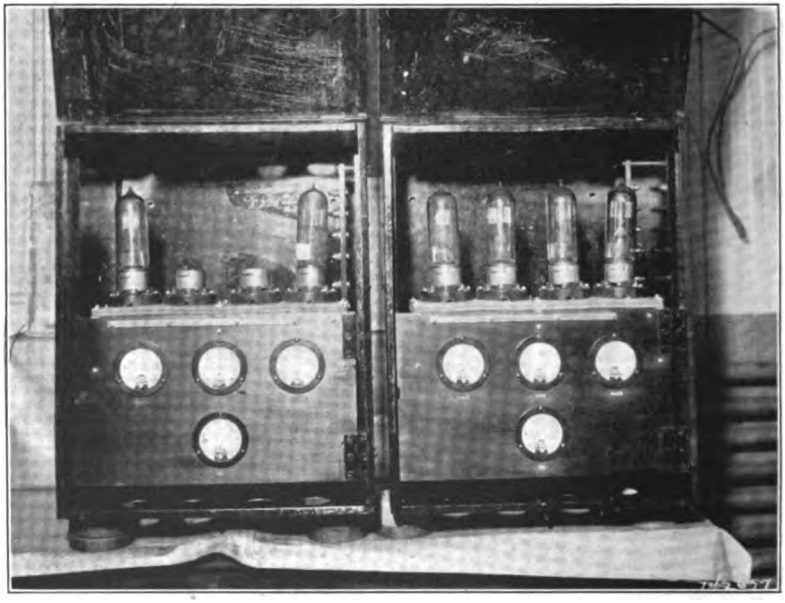

FIGURE 19—50-100-watt Speech Amplifier

STUDIO EQUIPMENT

Figure 20 shows the interior of the studio at the "Pittsburgh Post." Double carbon button and condenser type microphones are provided, each with its separate amplifier. Figures 21 and 22

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show the three-stage push-pull amplifier employed with the condenser microphone, and Figure 23 the control desk and amplifier for carbon microphones. A ventilated compartment contains

FIGURE 20—"Post" Studio, interior view

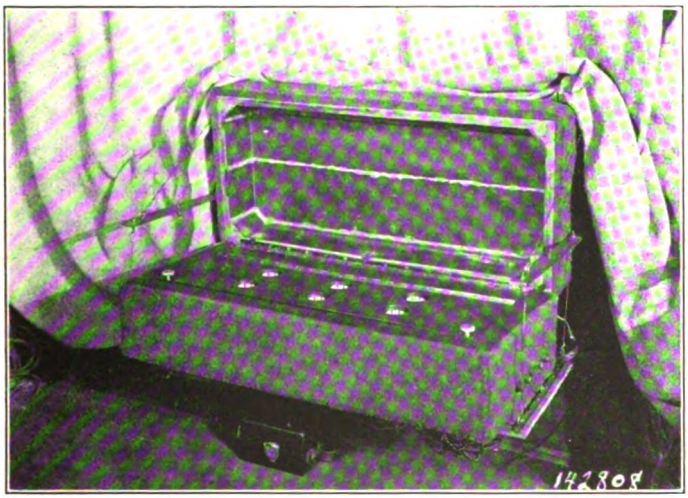

FIGURE 21—Three-stage Push-pull Amplifier, exterior

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duplicate 8-volt and 200-volt storage batteries for amplifier plate and filament power.

Figure 24 shows the East Pittsburgh studio where a glow microphone,[2] as well as condenser and double carbon microphones

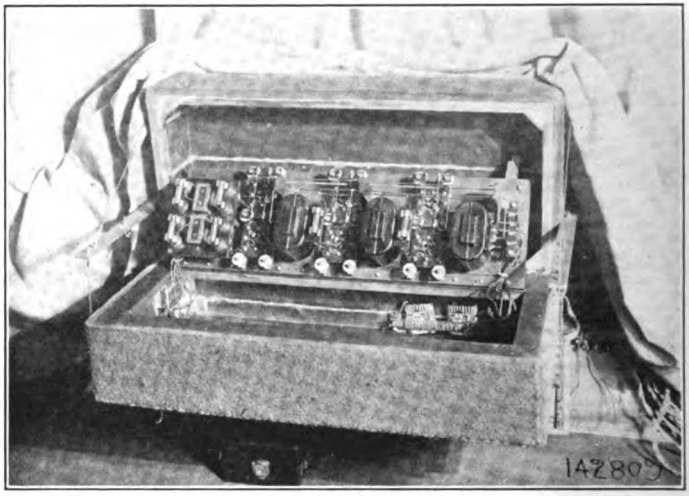

FIGURE 22—Three-stage Push-pull Amplifier, interior

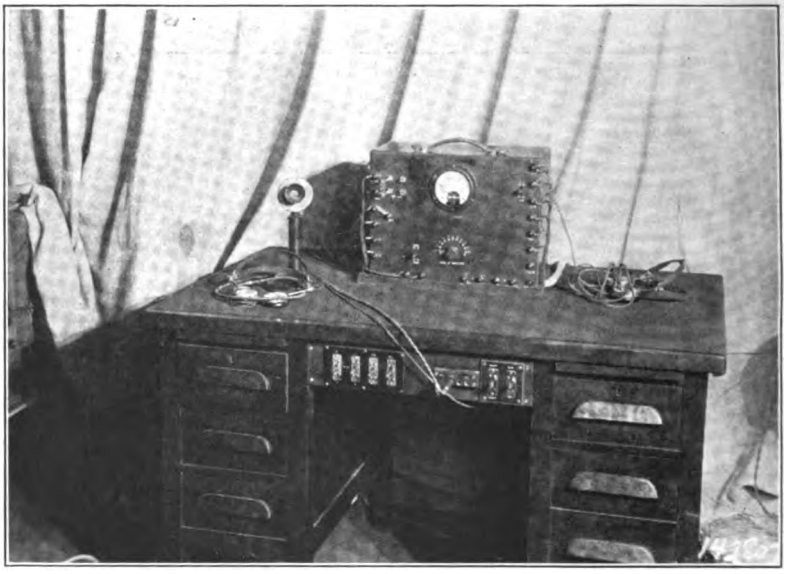

FIGURE 23—"Post" Studio Control Desk and Portable Amplifier

[2] For description, see "Journal of the American Institute of Electrical Engineers," volume XLII, number 3, page 219; March, 1922.

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are installed. The studio operator sits at a control desk, upon which are mounted the necessary amplifier change-over switches, signal lights, and modulation meter connecting with the station. There are 200-volt and 8-volt storage batteries with charging apparatus, which are permanently installed in an adjoining room.

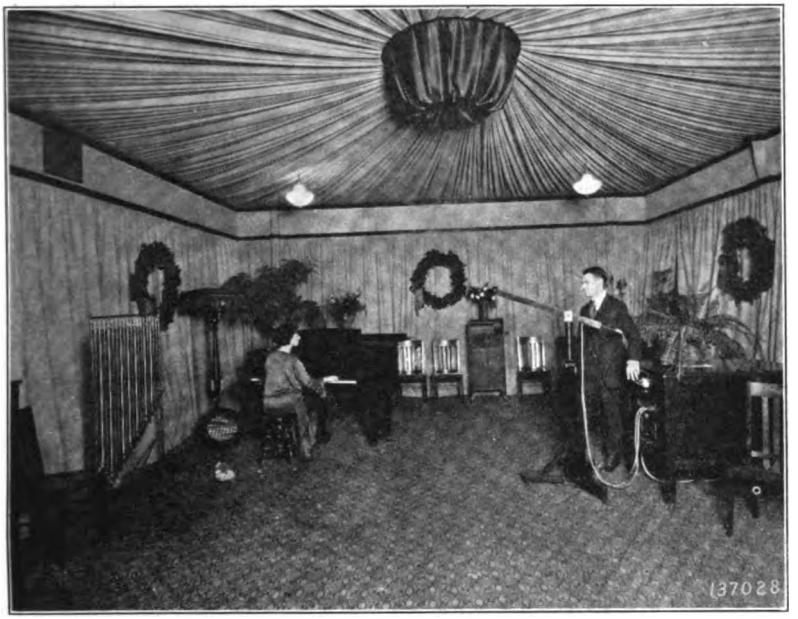

FIGURE 24—East Pittsburgh Studio, interior view

OUTSIDE PICK-UP APPARATUS

One type of portable amplifier employed at churches, ball parks, and other outside locations is shown on the desk in Figure 23. Double carbon button microphones are generally used, as many as eight being provided in the case of a large church. Figure 25 shows the amplifier and microphone switches at Calvary Church, Pittsburgh. An operator switches in the appropriate microphone for various parts of the service. For semi-permanent installations, such as the Grand Theatre, condenser microphones and suitable amplifiers are employed.

CONCLUSION

It is believed that KDKA, as described above, is at the present time second to no broadcasting station in the world. In point of area covered, as shown by letters and telegrams received,

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KDKA is regularly heard at night in all parts of the United States, in England, France, and Belgium. Several letters have also been received from South America and the Hawaiian Islands. The daylight range covers that part of the United States east of the Mississippi River.

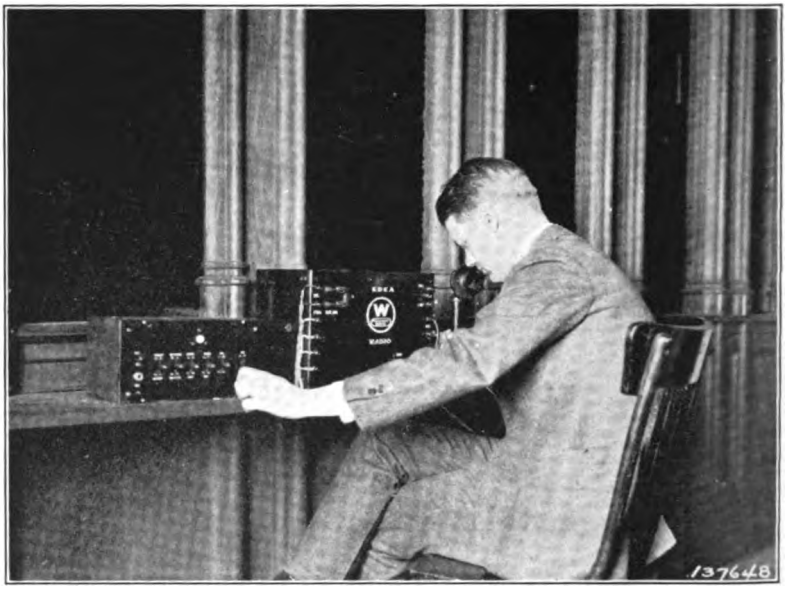

FIGURE 25—Amplifier and Microphone Control Switch at Calvary Church, Pittsburgh

The future of radio broadcasting as a public service depends to a large extent upon the quality of programs and of reproduction, together with the reliability of the transmitting stations. With the equipment described above it is believed that the service rendered by KDKA is a step in the right direction.

SUMMARY: The three transmitting sets used at East Pittsburgh are described in detail, together with their auxiliary apparatus.

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ON OPTIMUM HETERODYNE RECEPTION*

By

E. V. Appleton, M.A., D.Sc.,

and

Mary Taylor

(Cavendish Laboratory, Cambridge)

The reception of continuous-wave signals is nowadays carried out almost exclusively by the heterodyne method. In the ordinary applications of this method two oscillations are produced in the receiving assembly, one the result of the continuous-wave signal and the other by means of a local generator of constant frequency. The frequency of the local generator is so adjusted that the difference of local and received frequencies is well within audible range. If the receiving assembly were such that the effects of both local and received electromotive forces were strictly proportional to their magnitudes, that is, if the relations thruout were linear, the two oscillations would produce beats of audio frequency, but these would provoke no telephone response since the carrier wave would still be of radio frequency. It is only when the action of the receiving assembly is non-linear that the two oscillations give rise to a combination tone of frequency (n_1-n_2), equal to the difference of the two component frequencies n_1 and n_2. This combination tone gives rise to the telephonic signal. It may also be noticed, in passing, that the production of a combination tone of frequency (n_1-n_2) is necessarily accompanied by the production of a radio frequency combination tone of frequency (n_1+n_2).

Helmholtz's theory of combination tones is thus directly applicable to our problem with this difference that, whereas in the systems he considered, the non-linearity occurred in the restitution forces, in the heterodyne problems we are considering non-linearity is introduced by the use of resistances which do not obey Ohm's law. Resistances commonly used for this purpose are the contacts of dis-similar crystals or vacuum tube devices.

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In radio practice there are two methods of carrying out this production of combination tones. In one of these the generator producing the local oscillations is distinct from the detector circuit (separate heterodyne). In the other method the local oscillation is generated in the receiver itself and the non-linear properties of the latter are to a certain extent determined by the amplitude of this oscillation (autoheterodyne). Only the former of these methods will be considered here.

The practical advantages accruing from the use of the separate heterodyne method are well known, but to a certain extent practice has outdistanced theory. The chief relation of both practical and theoretical interest is the dependence of the amplitude of the combination tone on the amplitudes and frequencies of the component oscillations. The practical results in this connection were first obtained by E. H. Armstrong[1] whose experimental curves showed that the magnitude of the combination tone reached a maximum value as the electromotive force induced in the receiver from the local generator was increased. The ratio of the maximum amplitude of combination tone (optimum heterodyne) to that obtained when the received and local amplitudes were of equal magnitude (equal heterodyne) was found to depend on the signal strength, being sometimes as high as 55. Armstrong's results were thus at variance with the theoretical treatment due to Liebowitz,[2] whose work suggests that the greatest amplification obtainable by the use of the heterodyne method should be four times that obtained without heterodyne. The theory developed in the present paper shows that Armstrong's results are to be expected if the non-linear theory is properly applied and suggests other results which agree with data obtained experimentally.

The quantitative results of Armstrong[3] were obtained using cumulative grid rectification, but the optimum heterodyne phenomenon is to be found, tho sometimes not so marked, when anode rectification is used. We shall therefore deal with the quantitative aspects of both methods, but, since it happens that both the physical phenomena and the analysis are somewhat simpler in the case of anode rectification, we shall deal with that method first.

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[1] Proceedings of The Institute of Radio Engineers, volume 5, page 145, 1917.

[2] Proceedings of The Institute of Radio Engineers, volume 3, page 185, 1915.

[3] Previous citation.

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A—Anode Rectification

The most direct method of using the triode for the production of combination tones is that in which the electromotive forces due to the local and received oscillations are impressed between the grid and filament of the tube while the presence of the combination tones is indicated by the response of a telephone receiver T in the anode circuit. (See Figure 1.)

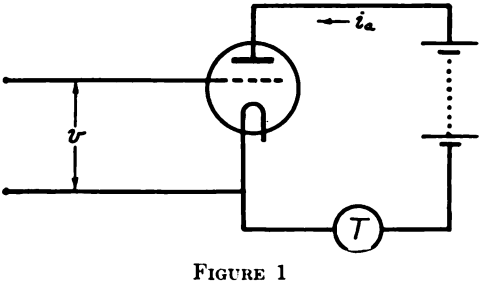

Figure 1

Recognizing the fact that combination tones in the telephone are only produced when the relation between the grid potential v and the anode current i_a is non-linear, let us assume that the relation between these quantities may be represented by a power series thus:

$$i_a = a_0 + a v + \beta v^2 + \gamma v^3 + \delta v^4 + \cdots. \tag{1}$$

Let us represent the magnitudes of the local and received oscillatory electromotive forces by $a \sin \omega_1 t$ and $b \sin \omega_2 t$, respectively. Thus if we choose as our origin of time the instant when the two oscillations are in phase we have

$$v = a \sin \omega_1 t + b \sin \omega_2 t \tag{2}$$

To find the amplitude A of the combination tone of angular frequency $(\omega_1 - \omega_2)$ we must substitute the expression for v from (2) in (1), and, after expanding the expression for i_a as a Fourier series pick out the terms involving $\sin (\omega_1 - \omega_2) t$ and $\cos (\omega_1 - \omega_2) t$. On doing this it is found that only the even terms (e.g., βv^2, δv^4, ζv^6, and so on) of (1) contribute to the expression for A. Thus if the power series is limited to the first three terms we have

$$A = \beta a b \tag{3}$$

while if it is limited to five terms we have

$$A = \beta a b + \frac{3}{2} \delta ab (a^2 + b^2) \tag{4}$$

and so on.

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If the triode is such that its characteristic may be represented by the first three terms of the power series we see from (3) that the stronger we make the local heterodyne oscillation the stronger becomes the combination tone, and thus no optimum effects are to be expected. With normal triodes, however, the electron emission from the filament is limited, and the anode current tends to saturation values as the grid potential is increased. Thus a three-term series is obviously inadequate. If a five-term series is used for i_a we find, on differentiating (4) with respect to a, that A passes thru a maximum value when

$$a^2 = a_o^2 = -\frac{2}{9}\frac{\beta}{\delta} - \frac{b^2}{3} \tag{5}$$

For the combination tone to pass thru a maximum value as the magnitude of the local oscillation is increased it is thus seen that β and δ must have opposite signs. An examination of the characteristics of typical triodes shows that this condition is normally fulfilled, β being positive and δ negative. Thus, in general, we see that for anode rectification there is an optimum value of the combination tone for a particular value of the electromotive force impressed on the receiver by the local oscillator and that this optimum local electromotive force depends on the characteristics of the triode and also, to a certain extent, on the magnitude of the signal electromotive force.

In view of what has just been said with regard to the sign of β and δ in (1) we shall, to avoid confusion, rewrite it as follows

$$i_a = a_o + \alpha v + \beta v^2 + \gamma v^3 - \delta v^4 + \cdots. \tag{1A}$$

The equation (1A) will be used in the remainder of the paper.

The optimum value A_o of the combination tone may be obtained by substituting for a^2 from (5) in (4), but the results are simpler if an approximation, consistent with the conditions of actual practice, is first made. Thus in a normal case the local oscillation amplitude a is usually large compared with the signal amplitude b and thus we may neglect b^2 in comparison with a^2. From (4) and (5) we find in this way

$$a_o^2 = \frac{2}{9}\frac{\beta}{\delta}, \tag{5A}$$

and

$$A_o = b\left(\beta a_o - \frac{3}{2}\delta a_o^3\right). \tag{6}$$

The practical significance of these results may be stated in words as follows:

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(5a) The magnitude of the local oscillation for optimum heterodyne is independent of the strength of the received signal if the latter is small. This is in agreement with Armstrong's experimental data.[4]

(6) The magnitude of the combination tone or signal is directly proportional to the strength of the received signal oscillation when the strength of the local oscillation has been adjusted for optimum heterodyne.

To get an idea of the magnitude of the quantities we are dealing with it is necessary to determine the values of β and δ in (1A) for a typical case. To determine these from the ordinary triode characteristics is a laborious matter, and a simpler alternative method has been devised. If we apply a sinusoidal electromotive force between the grid and filament of the tube and note the relation between the amplitude a of the applied electromotive force and the mean anode current $\bar{i}_a$ we obtain a characteristic from which the values of β and δ may be more directly deduced. For from equation (1A) we see that when v is equal to $a \sin \omega_1 t$ we have

$$\bar{i}_a = \frac{\beta a^2}{2} - \frac{3}{8}\delta a^4, \tag{7}$$

an expression not involving α, γ, ε, and so on. Such a mean current characteristic illustrating the relation between $\bar{i}_a$ and a for a triode is shown in Figure 2, from which it is obvious that β and δ as defined in (1A) are of the same sign. On the same diagram is shown an analogous characteristic for a crystal detector in which the mean current $\bar{i}$ thru the detector is shown as a function of the amplitude of the sinusoidal electromotive force across the detector. The characteristic is seen to be very similar to that of the triode. Both relations were determined experimentally, using low frequency electromotive forces derived from a resistance potentiometer.

The mean current characteristic of the triode is seen to have a point of maximum slope at Z. The appropriate value of a for this point may be found from (7) by differentiating twice with respect to a. Thus $\frac{d\bar{i}_a}{da}$ is a maximum when

$$a^2 = \frac{2}{9}\frac{\beta}{\delta}$$

We note, however, from (5A) that this is the value of a for

[4]Armstrong, previous citation, Figure 4, page 152. See also Austin and Grimes, "Wash. Acad. Sci.," March 19, 1920.

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optimum heterodyne and we thus see that when the received signal is very small the value of the local oscillation amplitude should be adjusted to the value represented by OX, where the slope of the mean anode current characteristic is a maximum.

In a case where the received signal is not small the more accurate formula must be used, the optimum value of a being given by (5), and the corresponding amplitude of the combination tone by (4).

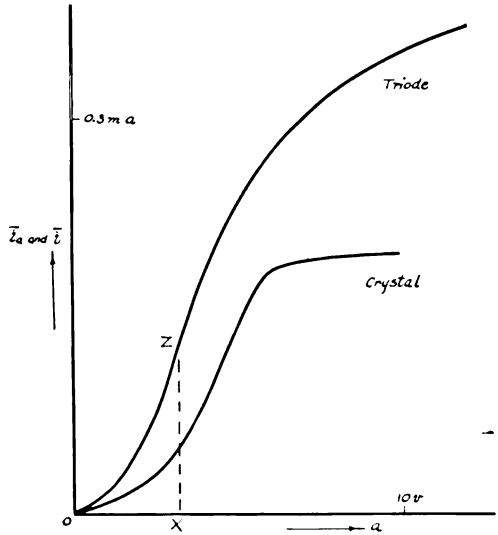

FIGURE 2

It is a simple matter now to state exactly how sensitive heterodyne reception really is. For example, if the signal is small, we see that the current of the combination tone deduced from (5A) and (6) is given by

$$A_o = \frac{2}{3}\beta a_o b,$$

or

$$A_o = \frac{2\sqrt{2}}{9}\beta^{\frac{1}{2}}\delta^{\frac{1}{2}} \cdot b \qquad (9)$$

On substituting the appropriate values of β and δ for the characteristic of Figure 2 (triode) we find that

$$A_o = 6.5 \times 10^{-2} \ m.a. \text{ per signal volt.}$$

To illustrate the extraordinary sensitivity of the heterodyne method when the local oscillation amplitude is adjusted for

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optimum conditions, we may compare the optimum value of the combination tone with that obtained in the case of equal heterodyne when a is equal to b. It will be seen later that such a comparison is in effect a true comparison of the sensitivity of a triode receiver for continuous-wave signals and spark (or completely modulated wave) signals, since the signal strength obtained with equal heterodyne is equal to that obtained with spark or modulated wave signals of the same intensity. From (4) we see that the amplitude of the combination tone (A_E) for equal heterodyne is given by

$$A_E = \beta b^2 - 3\delta b^4$$

or, when b^2 is small by

$$A_E = \beta b^2$$

We thus find that

$$\frac{A_o}{A_E} = \frac{2\sqrt{2}}{9}\sqrt{\frac{\beta}{\delta}} \cdot \frac{1}{b}, \qquad (11)$$

from which it is seen that the advantages accruing from the use of the optimum heterodyne method are more marked the smaller the received signal.[5] To give definiteness to this statement we have calculated the ratio of the optimum heterodyne signal to that obtained with equal heterodyne (or spark or modulated wave signal) for a triode and various strengths of received signal. These are tabulated below.

b (volt)	$\frac{A_o}{A_E}$	Formula
$\frac{1}{10}$	21	
$\frac{1}{50}$	105	$\frac{A_o}{A_E} = \frac{2\sqrt{2}}{9}\sqrt{\frac{\beta}{\delta}} \cdot \frac{1}{b}$
$\frac{1}{200}$	420	where $\beta = 0.0308 \frac{m.a.}{(volts)^2}$
$\frac{1}{1000}$	2100	$\delta = 0.00069 \frac{m.a.}{(volts)^4}$
$\frac{1}{5000}$	10200	

[5] Compare Armstrong's experimental results, previous citation, Figure 2, page 149.

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The amplification obtainable by the use of optimum heterodyne may thus be enormous.

B—Cumulative Grid Rectification

As this method of rectification has not previously been the subject of a complete quantitative investigation we propose to deal with the matter fairly thoroly. The main experimental features of this method are that a condenser of small capacity is inserted in the grid circuit of a triode assembly while a high resistance leak is connected in parallel with the grid-filament path. (See Figure 3.) High frequency (radio frequency) potential changes are applied between the points A and B while the resulting low frequency (audio frequency) effect is detected in the anode circuit. The low-frequency effect may take the form of a diminution of the mean anode current, in which case some form of automatic recorder is used, or it may take the form of the production of combination tones by a telephone inserted in the anode circuit.

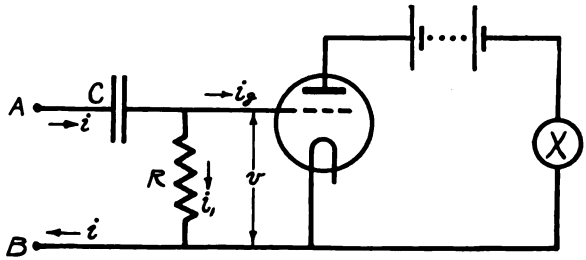

Figure 3—Showing Essential Features of Cumulative Grid Rectification Circuit. X is Automatic Recorder or Telephone

Let us denote the high-frequency electromotive force applied between A and B as E, and let the grid potential and grid current be denoted by v and i_g, respectively. Let also the current thru the condenser C be denoted by i and that thru the resistance by i_1. The expression of Kirchhoff's laws is therefore

$$v = R\,i_1 = E - \int \frac{i}{C}\,dt, \tag{12}$$

and

$$i = i_1 + i_g \tag{13}$$

Let us assume that the relation between the grid current and grid potential is not linear but may conveniently be expressed in the form of a power series thus

$$i_g = f(v) = \alpha_1 v + \beta_1 v^2 + \gamma_1 v^3 + \cdots. \tag{14}$$

where α_1, β_1, γ_1, and so on, are constants for the particular

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triode, filament temperature, and anode voltage used. It may here be noted that we are neglecting the small and constant current which flows thru the grid circuit and the resistance R, and which would be represented by a constant term in the expression for i_g. If, however, we take as the zeros of current and potential those existing in the absence of signals, the expression (14) may be regarded as generally valid.

The result of eliminating i, i_1, and i_g from (12) and (13) is

$$\frac{dv}{dt}+\frac{f(v)}{C}+\frac{v}{CR}=\frac{dE}{dt}, \tag{15}$$

a differential equation of a generalized Riccati type which represents the action of the grid condenser. The approximate solution of this equation when E has different forms, representing different types of impressed signal, must be investigated. For examples we shall consider the following cases of practical importance.

(1) The reception of continuous waves without heterodyne. In this case the impressed force between the points A and B is sinusoidal and we thus write

$$E=b\sin\omega t \tag{16}$$

(2) The reception of modulated continuous waves. In this case the alternating electromotive force of angular frequency ω is modulated with an angular frequency p and we write

$$E=b(1+\sin pt)\sin\omega t \tag{17}$$

(3) The reception of continuous waves by the heterodyne method. In this case the impressed electromotive force consists of a received signal of angular frequency ω_2 and a constant and local oscillation of angular frequency ω_1. Thus

$$E=a\sin\omega_1 t+b\sin\omega_2 t \tag{18}$$

We shall consider these three cases in the order mentioned, and, as a first approximation, we shall limit the power series of (14) to the first two terms which are sufficient to enable us to explain the outstanding physical phenomena.

Reception of Continuous Waves Without Heterodyne

In this case the representative differential equation becomes (See (14), (15), and (16))

$$\frac{dv}{dt}+\left(\frac{1}{RC}+\frac{\alpha_1}{C}\right)v+\frac{\beta_1 v^2}{C}=b\omega\cos\omega t \tag{19}$$

Let us suppose, as is usually the case in practice, that the

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term involving β is small, and that, as a first approximation, it may be neglected. Omitting this term the solution of (19) for the steady state is

$$v = \frac{b\omega}{\sqrt{\omega^2 + \frac{q^2}{C^2}}} \cos(\omega t - \phi)$$

where $\tan \phi = \frac{\omega C}{q}$, and $q = \frac{1}{R} + \alpha_1$.

On substituting this value for v in the non-linear term involving β_1 in (19) the approximate representative equation takes the form

$$\frac{dv}{dt} + \frac{q}{C} v + \frac{\beta_1}{2C} \left(\frac{b^2 \omega^2}{\omega^2 + \frac{q^2}{C^2}} \right) \{1 + \cos 2(\omega t - \phi)\} = b\omega \cos \omega t. \quad (20)$$

The solution of (20) for the steady state consists of three parts

(1) a constant term $v = -\frac{\beta_1}{2q} \left(\frac{b^2 \omega^2}{\omega^2 + \frac{q^2}{C^2}} \right)$,

(2) a term of angular frequency ω given by

$$v = \frac{b\omega}{\sqrt{\omega^2 + \frac{q^2}{C^2}}} \cos(\omega t - \phi),$$

and (3) a term of angular frequency 2ω the magnitude of which is proportional to β_1.

The first of these is of special interest, for it shows that, when a sinusoidal electromotive force is applied to the assembly, the mean grid potential is reduced by an amount proportional to the square of the amplitude of the applied electromotive force. This reduction of mean grid potential produces a diminution of mean anode current by means of which the presence of the high frequency signal is detected or measured.[6]

Reception of Completely Modulated Waves

In this case the representative equation is (See (14), (15), and (17))

$$\frac{dv}{dt} + \frac{q}{C} v + \frac{\beta_1}{C} v^2 = b \left[\omega \cos \omega t - \frac{\omega - p}{2} \sin(\omega - p)t + \frac{\omega + p}{2} \sin(\omega + p)t \right] \quad (21)$$

[6] Compare Moullin, "Wireless World," volume 10, page 1, 1922.

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Since the term involving β_1 is small the first approximation to the solution of (21) is given by the solution of

$$\frac{dv}{dt}+\frac{q}{C}v=b\left[\omega\cos\omega t-\frac{\omega-p}{2}\sin(\omega-p)t+\frac{\omega+p}{2}\sin(\omega+p)t\right]. \quad (21A)$$

For the steady state this solution is

$$v=\frac{b\omega C}{\sqrt{q^2+C^2\omega^2}}\cos\left(\omega t-\tan^{-1}\frac{\omega C}{q}\right)-\frac{bC(\omega-p)}{2}\frac{\sin\left[(\omega-p)t-\tan^{-1}\frac{C(\omega-p)}{q}\right]}{\sqrt{q^2+C^2(\omega-p)^2}}+\frac{bC(\omega+p)}{2}\frac{\sin\left[(\omega+p)t-\tan^{-1}\frac{C(\omega+p)}{q}\right]}{\sqrt{q^2+C^2(\omega+p)^2}} \quad (22)$$

But in actual practice, as will be shown later, we have

$$C\omega\gg q$$

and

$$\omega\gg p$$

Thus (22) may be written approximately

$$v=b\left[\sin\omega t+\frac{1}{2}\cos(\omega-p)t-\frac{1}{2}\cos(\omega+p)t\right] \quad (23)$$

On substituting for v from (23) in the term involving β_1 in (21) we obtain

$$\frac{dv}{dt}+\frac{q}{C}v+\frac{\beta_1}{C}b^2\left[\frac{3}{4}-\frac{1}{2}\cos 2\omega t+\frac{1}{8}\cos 2(\omega-p)t+\frac{1}{8}\cos 2(\omega+p)t+\sin pt+\frac{1}{2}\sin(2\omega-p)t-\frac{1}{2}\sin(2\omega+p)t-\frac{1}{4}\cos 2\omega t-\frac{1}{4}\cos 2pt\right]=b\left[\omega\cos\omega t-\frac{\omega-p}{2}\sin(\omega-p)t+\frac{\omega+p}{2}\sin(\omega+p)t\right] \quad (24)$$

From this we may see that the variation in the grid potential of signal frequency, which is the point of special interest, is found from the equation

$$\frac{dv}{dt}+\frac{q}{C}v=-\frac{\beta_1 b^2}{C}\sin pt \quad (24A)$$

Its amplitude A is therefore given by

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$$A = \frac{\beta_1 b^2}{\sqrt{q^2 + C^2 p^2}} \qquad (25)$$

Thus we see that, due to the action of the grid condenser, a completely modulated wave gives rise to a low-frequency grid potential variation of amplitude given by (25) which, causing corresponding changes of anode current, gives rise to telephonic response. This amplitude is equal to that obtained with equal heterodyne. (See (29) and (29A) later.)

Reception of Continuous Waves by Heterodyne Method

In this case the representative equation becomes (see (14), (15), and (18))

$$\frac{dv}{dt} + \frac{q}{C} v + \frac{\beta_1}{C} v^2 = a\,\omega_1 \cos \omega_1 t + b\,\omega_2 \cos \omega_2 t \qquad (26)$$

If again we omit the β_1 term a first approximation of (26) is given by the solution of

$$\frac{dv}{dt} + \frac{q}{C} v = a\,\omega_1 \cos \omega_1 t + b\,\omega_2 \cos \omega_2 t. \qquad (26A)$$

For the steady state this is

$$v = \frac{a\,\omega_1}{\sqrt{\omega_1^2 + \frac{q^2}{C^2}}} \cos(\omega_1 t - \phi_1) + \frac{b\,\omega_2}{\sqrt{\omega_2^2 + \frac{q^2}{C^2}}} \cos(\omega_2 t - \phi_2)$$

where $\tan \phi_1 = \frac{C\,\omega_1}{q}$ and $\tan \phi_2 = \frac{C\,\omega_2}{q}$.

We may write this

$$v = r \cos(\omega_1 t - \phi_1) + s \cos(\omega_2 t - \phi_2)$$

where $r = \frac{a\,\omega_1}{\sqrt{\omega_1^2 + \frac{q^2}{C^2}}}$ and $s = \frac{b\,\omega_2}{\sqrt{\omega_2^2 + \frac{q^2}{C^2}}}$.

Thus for substitution in the original equation we have

$$v^2 = \frac{1}{2}\Big[r^2 + s^2 + r^2 \cos 2(\omega_1 t - \phi_1) + s^2 \cos 2(\omega_2 t - \phi_2) + 2\,r\,s \cos\{(\omega_1 - \omega_2)t + \phi_1 - \phi_2\} + 2\,r\,s \cos\{(\omega_1 + \omega_2)t - \phi_1 - \phi_2\}\Big]$$

The solution of (26) when the substitution for v^2 has been made will be of the form

$$v = M + r \cos(\omega_1 t - \phi_1) + s \cos(\omega_2 t - \phi_2) + B \cos 2(\omega_1 t - \phi_1') + C \cos 2(\omega_2 t - \phi_2') + D \cos\{(\omega_1 - \omega_2)t + \phi\} + E \cos\{(\omega_1 + \omega_2)t + \psi\} \qquad (27)$$

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where A, B, C, D, ϕ_1', ϕ_2', ϕ, and Ψ are constants depending on the constants of the circuit and on a, b, ω_1, and ω_2.

For practical purposes interest centers in the constant term M which represents the alteration of mean grid potential, and in the coefficient D which gives the amplitude of the grid potential variation of the signal angular frequency $(\omega_1-\omega_2)$. These are found to be as follows

$$M=-\frac{\beta_1}{2q}(r^2+s^2)=-\frac{\beta_1}{2q}\left(\frac{a^2C^2\omega_1^2}{q^2+C^2\omega_1^2}+\frac{b^2C^2\omega_2^2}{q^2+C^2\omega_2^2}\right) \qquad (28)$$

and $$D=\frac{\beta_1 r s}{\sqrt{q^2+C^2(\omega_1-\omega_2)^2}}$$

$$=-\beta_1 a b\frac{C^2\omega_1\omega_2}{\sqrt{q^2+C^2(\omega_1-\omega_2)^2}\sqrt{q^2+C^2\omega_1^2}\sqrt{q^2+C^2\omega_2^2}} \qquad (29)$$

The presence of the received amplitude b may therefore be detected as an alteration of mean anode current produced by the alteration of mean grid potential M, or by an alteration of anode current of angular frequency $(\omega_1-\omega_2)$, caused by the low frequency variation of grid potential D. It is the latter variation of anode current which gives rise to the well-known heterodyne note in the anode circuit telephone.

In considering the physical processes underlying this method of producing combination tones we note that the grid condenser performs two functions. In the first place it hands on the potential changes to the grid, while in the second place it effects a partial insulation of the grid so that the potential across it is sensibly the same as the grid potential itself. So far as the first function is concerned it is advantageous to have the grid condenser as large as possible, but consideration of the second function shows that a very large capacity would tend to smooth out the grid potential low-frequency changes. There must, therefore, be an optimum grid condenser value determined by the frequency of the oscillations ($2\pi\omega_1$ or $2\pi\omega_2$), by the combination tone frequency ($2\pi(\omega_1-\omega_2)$) and by the triode and circuit parameters. This value is found by differentiating the expression for D with respect to C, the result being that[7]

$$C_{opt}^2=\frac{\sqrt{2}\,q^2}{\omega(\omega_1-\omega_2)}$$

It is of interest to compare this optimum value of C with the value usually used in practice. As typical practical values we may use

$$q=1.8\times10^{-5}\,(ohms)^{-1}$$
$$\omega=10^6\,(secs.)^{-1}$$

[7] We have here assumed that $\omega_1\doteq\omega_2=\omega$.

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and $$\omega_1 - \omega_2 = 2\pi\,(1{,}000)\ (secs.)^{-1}$$

The theoretical value of $C_{opt.}$ is thus 0.003 mfds., which is in close agreement with the value normally used.

We may further note the following useful approximations.[8] Since it is usual to work with the optimum condenser value we have

$$C^2 = \frac{\sqrt{2}\,q^2}{\omega\,(\omega_1 - \omega_2)}$$

Also $$\frac{\omega_1 - \omega_2}{\omega} \ll 1$$

and thus $$C\,\omega \gg q$$

and $$C\,(\omega_1 - \omega_2) \ll q.$$

Using the above approximations (28) and (29) may be written

$$M = -\frac{1}{2}\frac{\beta_1}{q}(a^2 + b^2), \tag{28A}$$

and $$D = -\frac{\beta_1}{q}\,a\,b. \tag{29A}$$

We thus see that as a first approximation the amplitude of the low-frequency grid potential changes produced by grid condenser action is proportional to the product of the received and local signal amplitudes. If the relation between the grid potential and the anode current is linear, the amplitude of the combination tone produced is also proportional to the same product. This result may be compared with the first approximation made in the case of anode rectification, where a similar result was obtained. (See (3)). A further investigation (the algebra of which is too lengthy to be given here) shows that the parallelism between anode rectification and cumulative rectification is quite complete, in that, if the power series (14) for i_g contains a δ term, an expression of a type similar to (4) is obtained. Thus optimum heterodyne effects are possible with cumulative grid rectification if the power series for i_g contain a negative δ term.

But when the relation between the grid potential and anode current is not linear, as is often the case, optimum heterodyne effects are obtainable which have no parallel in the case of anode rectification, and we now proceed to discuss such cases in greater detail.

Let us assume that the relation between the anode current and grid potential may be written as before

$$i_a = \alpha\,v + \beta\,v^2, \tag{1}$$

[8] Some of these approximations have been previously used. See page 13

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where the zeros of potential and current are the values obtaining in the absence of local and distant signals. In the presence of such signals the value of v is given by (27), but so far as the production of combination tones is concerned the important terms are given by[9]

$$v = M + a \sin \omega_1 t + b \sin \omega_2 t + D \cos \{(\omega_1 - \omega_2) t\},$$

where M and D are given by (28A) and (29A).

On substituting this value for v in (1) and collecting terms of the angular frequency $(\omega_1 - \omega_2)$ the amplitude A of the combination tone is found to be given by

$$A = a b \left[\beta - \frac{a \beta_1}{q} + \frac{\beta \beta_1^2}{q^2} (a^2 + b^2)\right]. \qquad (30)$$

Thus when the value of β is positive the amplitude of the combination tone of angular frequency $(\omega_1 - \omega_2)$ passes thru a maximum value as a the amplitude of the local oscillation is increased. The optimum value of a is given by

$$a^2 = \frac{1}{3}\left[\frac{q(a\beta_1 - q\beta)}{\beta \beta_1^2} - b^2\right]$$

In practice the values of b are very small compared with those of a. The optimum value of a is thus independent of b and given by

$$a^2 = \frac{1}{3}\left[\frac{q(a\beta_1 - q\beta)}{\beta \beta_1^2}\right] \qquad (31)$$

The value of β is usually positive in practice when low anode potentials are used, so that optimum heterodyne phenomena due to the combined non-linear properties of the grid and anode circuits are obtained. With larger values of anode potential, however, the value of β is sometimes negative in which case the above considerations do not apply. For such conditions the effects of anode rectification and cumulative grid rectification are additive and not in opposition. The study of such cases would most probably be profitable from a practical point of view.

To illustrate the optimum heterodyne phenomena mentioned above the mean-current characteristic representative of grid condenser action in a typical case is shown in Figure 4. This diagram shows the variation of the mean anode current $\bar{i}_a$ as a funtion of the amplitude of a sinusoidal electromotive force $(E = a \sin \omega_1 t)$ and may be compared with Figure 2 which is the equivalent characteristic for anode rectification.

[9] The approximations previously mentioned (page 17) have here been used.

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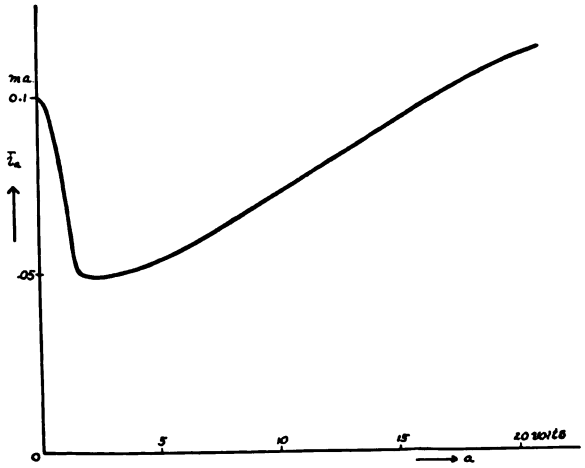

FIGURE 4

In this case the solution of (20) for the appropriate value of E is

$$v = -\frac{\beta_1 a^2}{2q} + a\cos(\omega_1 t - \phi) + F\cos(2\omega_1 t - \chi), \qquad (32)$$

where F is proportional to ω_1^2 and $C\omega_1 \gg q$. Substituting this expression for v in (1) the variation of the mean anode current $\bar{i}_a$ becomes

$$\bar{i}_a = -\frac{\alpha\beta_1}{2q}a^2 + \beta\left[\frac{\beta_1^2}{4q^2}a^4 + \frac{1}{2}(a^2+F^2)\right] \qquad (33)$$

Here F^2 may be neglected since it is of the order β_1^4. The value of $\bar{i}_a$ has a turning point when

$$a^2 = \frac{q^2}{\beta\beta_1^2}\left(\frac{\alpha\beta_1}{q} - \beta\right). \qquad (34)$$

For real values of a it is, therefore, necessary that $\alpha\beta_1 > q\beta$, and in this case the mean anode current passes thru a minimum for increasing a; it is initially diminished.

The value of a at the turning point depends on the triode characteristics and the value of the grid leak R. In the actual case represented in Figure 4 the values of the triode parameters were independently found to be

$$\alpha_1 = 0.017\,\frac{m.a.}{volt}; \qquad \beta_1 = 0.013\,\frac{m.a.}{(volts)^2}$$

$$\alpha = 0.013\,\frac{m.a.}{volt}; \qquad \beta = 0.024\,\frac{m.a.}{(volts)^2}$$

$$R = 10^6\ ohms, \text{ and thus } q = 0.018\,\frac{m.a.}{volt}.$$

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Substituting these values in (34) theory indicates that the value of a at the turning point should be 2.02 volts; the experimental value was 2.05 volts.

The slope of the $(a, \bar{i}_a)$ curve is a maximum when

$$a^2 = \frac{1}{3} \frac{q^2}{\beta \beta_1^2} \left(\frac{\alpha \beta_1}{q} - \beta \right) \tag{35}$$

This is the same value of a as that given by (31) for the value of the local amplitude when optimum heterodyne effects occur. We thus see that the appropriate value of a for optimum heterodyne reception may be found by drawing a curve such as that of Figure 4 for the particular triode in question and finding the value of the applied emf. at the point of maximum slope. Zero combination tone effects are obtained if a has the value given by (34) for the turning point of the curve.

It will be further noticed that for continuous wave reception without heterodyne the maximum anode current response is given by a signal of strength given by (34) and that for a signal very much stronger than this the response might be negligible.

In conclusion we should like to express our gratitude to Dr. B. van der Pol for his kind interest in this work.

SUMMARY: The use of the triode for anode rectification and cumulative grid rectification in heterodyne reception is investigated theoretically, and it is shown that:—

(a) The amplitude of the combination tone produced in a radio receiver telephone reaches an optimum value for a certain value of the local oscillation amplitude. In anode rectification this phenomenon is due to the non-linear nature of the grid potential/anode current relation, while in grid condenser action the phenomenon may be due either to the non-uniform grid conductance or to the combined non-linearity of the relations between grid potential and grid and anode currents.

(b) The optimum value of the local oscillation amplitude is independent of the received signal magnitude when the latter is small.

(c) The magnitude of the combination tone is directly proportional to the strength of the signal oscillation when the local oscillation has been adjusted for optimum heterodyne.

(d) The ratio of the optimum value of the combination tone to that produced with equal heterodyne is inversely proportional to the strength of the received signal.

(e) The differential equation representative of grid condenser action is shown to be of a generalized Riccati type.

$$\frac{dv}{dt} + f(v) = \phi(t).$$

Approximate solutions of this are obtained for practical cases. The optimum grid condenser value obtained theoretically in this way is in good agreement with the practical value.

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SIGNAL-TO-STATIC INTERFERENCE RATIO IN RADIO TELEPHONY*

By

Marius Latour

(Paris, France)

Having read Dr. J. R. Carson's communication which appeared in the June, 1923, issue of the Proceedings of The Institute of Radio Engineers, I desire to add the following material:

I have previously proposed to carry out radio telephony without carrier transmission and thereby implying homodyne reception[1] But I believe that I have since then found an arrangement which is paramount from the all-important point of view of signal-to-static ratio, and which might advantageously be used in case of shorter waves where homodyne reception is practically impossible, and where the speech frequency represents a relatively much smaller fraction of the wave length spectrum.

The speech voltages occurring at the secondary terminals of a telephone induction coil being of the shape represented by the curve of Figure 1, a radio telephone transmitter antenna is traversed by a radio-frequency current which, whilst having an amplitude which is constantly modulated proportionally to the speech voltages, is of a certain frequency f_1 during positive speech impulses and of a different frequency f_2 during negative speech impulses. These results may be arrived at by means of

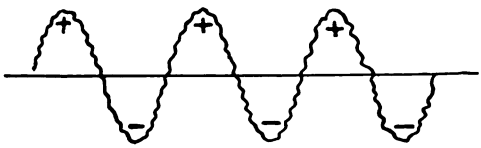

Figure 1

the arrangement represented schematically in Figure 2. The primary current variations produced by the action of the micro-

* Received by the Editor, August 3, 1923.

[1] See my French patent, number 21.853/512,395 of 1916.

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phone 1 are amplified by the vacuum tube amplifier 2, which comprises an input transformer 3 and an output transformer 4. The speech voltages thus amplified are used to supply the generating tubes 5 and 6 according to the arrangement of Figure 2, aiming at straight alternating current supply of vacuum tubes.[2] Tubes 5 and 6 work alternately. The waves successively emitted by each tube may be made to have different lengths by different grid or plate circuit settings which we will not here describe.

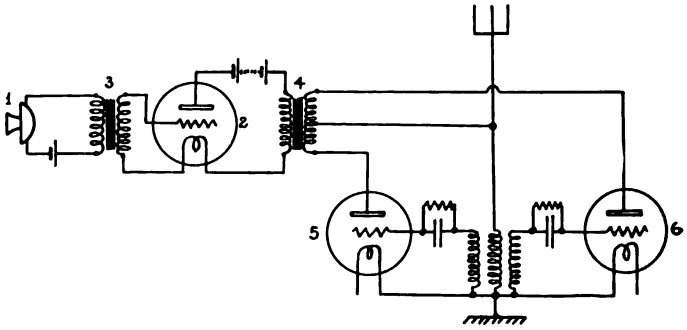

FIGURE 2

Instead of sending the radio-frequency current directly into the antenna, we may, of course, insert a radio-frequency power amplifier between the arrangement of Figure 2 and the antenna.

With the above system of transmission, the reception could be carried out in accordance with the arrangement shown in Figure 3, which is in itself a means of reducing static interference.[3]

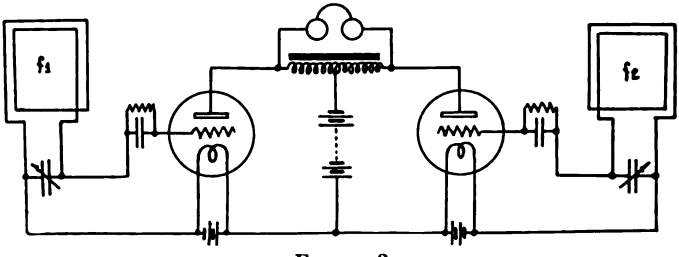

FIGURE 3

It may be seen that, from the point of view of detection of static-provoking shock excitation of the receiving circuits, the two circuits being tuned to different wave-lengths are in opposi-

[2] This is the object of my French patent numbers 502, 601 of 1915

[3] See my French patent, number 526,674.

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tion to each other, while, from the point of view of speech reception, the two circuits work alternately but concurrently.

The power supplied to the antenna is not higher than that in carrierless transmission using homodyne reception.

From the point of static, it must be borne in mind that the introduction of a carrier-wave either at the transmitting end or at the receiving end, which is inherent to ordinary radio telephony, is apt to double the intensity of static currents susceptible of having the same frequency as the carrier wave and occurring sometimes in phase and at other times in opposition to the latter.

The method just described of transmitting and receiving without carrier wave does not possess this drawback.

Paris, July 24th, 1923.

SUMMARY: Referring to Dr. Carson's paper on "Static-to-Signal Ratio," the author discloses a system of twin side-band radiotelephonic transmission, wherein the carrier component is suppressed, but which differs from other carrierless transmitting systems in that a homodyne receiver is not required. This last peculiarity makes the system particularly suitable for short wave-length transmission and especially so in view of the fact that the disadvantage of broader wave-length spectrum inherent to twin side-band transmission vanishes as the wave-length decreases.

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REGENERATION IN COUPLED CIRCUITS*

By

E. Leon Chaffee

(Associate Professor of Physics, Harvard University, Cambridge, Massachusetts)

Part I

Single Circuit Regeneration

It is well known that a vacuum tube, regeneratively connected to a single circuit, acts to reduce the effective resistance of that circuit. At the same time the effective reactance is usually altered. This reduction in effective resistance is often spoken of as due to the introduction of a negative resistance by the tube. This effective negative resistance may or may not be a function of the frequency. When the total effective resistance is equal to or less than zero the circuit oscillates.

The analytical discussion of single-circuit regeneration is given below. There are two cases according to whether the oscillating element is in the plate circuit or in the grid circuit of the tube.

Case I. Oscillating Element in Plate Circuit. In Figure 1, let L, C, and R be the constants of the oscillatory circuit, L_g the grid circuit inductance and m the mutual inductance between L and L_g. Let $R_p = \frac{\partial e_p}{\partial i_p}$ be the plate to filament resistance and $\mu = -\frac{\partial e_p}{\partial e_g}$ be the amplification factor of the tube. The currents are indicated in Figure 1. Let E be the effective value of an emf. of frequency $\frac{\omega}{2\pi}$ induced in L from an external source. The following complex equations express the conditions in Figure 1. The grid current is assumed to be negligible.

Quantities in Clarendon Type are complex.

*Received by the Editor, February 15, 1924.

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$$\begin{cases} (R + jL\omega) I_1 - \frac{j}{C\omega} I_2 = E \\ \frac{j}{C\omega} I_2 + R_p I_p = \mu E_g = j\mu m \omega I_1 \\ I_1 = I_p + I_2 \end{cases}$$

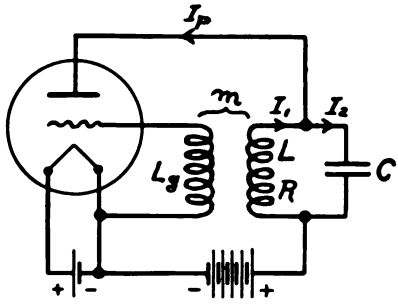

FIGURE 1

The solution of these equations is

$$I_1 = \frac{E}{\left[R + \frac{\frac{1}{C^2\omega^2} - \frac{\mu m}{C}}{R_p^2 + \frac{1}{C^2\omega^2}} R_p\right] + j\left[L\omega - \frac{1}{C\omega} \cdot \frac{R_p^2 + \frac{\mu m}{C}}{R_p^2 + \frac{1}{C^2\omega^2}}\right]} \tag{1-1}$$

An examination of equation (1-1) shows that the effective or equivalent resistance is decreased, due to the term $\frac{\mu m}{C}$, and will be zero if

$$\frac{\mu m R_p}{C} = R\left(R_p^2 + \frac{1}{C^2\omega^2}\right) + \frac{R_p}{C^2\omega^2} \tag{1-2}$$

at which point oscillation commences. The effective capacity reactance is increased as shown by the imaginary part of equation (1-1).

Equation (1) and subsequent equations are much more general and hence of greater use if all constants of the circuits are reduced to coefficients and all angular velocities or wave lengths to ratios of angular velocities or wave lengths. By this reduction the equations can be more easily applied to any circuit. To this end the following definitions are made.

Let $\frac{\omega_o}{2\pi}$ be a particular frequency to which any frequency $\frac{\omega}{2\pi}$ may be referred in the form of a ratio $\frac{\omega_o}{\omega}$. The ratio $\frac{\omega_o}{\omega} = \frac{\lambda}{\lambda_o}$ will

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be denoted by θ so that for ω in all equations $\frac{\omega_o}{\theta}$ may be substituted.

If $\frac{\omega_1}{2\pi}$ and $\frac{\omega_2}{2\pi}$ represent the natural undamped frequencies of oscillation of two circuits, then $\frac{\omega_o}{\omega_1}$ will be denoted by θ_1 and $\frac{\omega_o}{\omega_2}$ by θ_2, respectively. It is then obvious that

$$\theta_1 = \frac{\lambda_1}{\lambda_o} = \frac{\omega_o}{\omega_1} = \omega_o \sqrt{L_1 C_1} \text{ and } \theta_2 = \frac{\lambda_2}{\lambda_o} = \frac{\omega_o}{\omega_2} = \omega_o \sqrt{L_2 C_2}.$$

The following coefficients are here defined; others will be introduced later when required.

$$\frac{R}{L\omega_o} = \eta, \frac{R_p}{L\omega_o} = \eta_p, \sqrt{\frac{m}{L L_p}} = K, \text{ or } \frac{m}{\sqrt{L L_g}} = K$$

Using the above ratios and coefficients, equation (1-1) reduces to (1-a)

$$\text{(1-a)} \qquad I_1 = \frac{E}{L\omega_o \left[\eta - \frac{\mu K \theta_1^2 \sqrt{\frac{L_g}{L}} - \theta^2}{\eta_p^2 \theta_1^4 + \theta^2} \eta_p\right] + \frac{j}{\theta}\left[1 - \theta^2 \frac{\eta_p^2 \theta_1^2 + \mu K \sqrt{\frac{L_g}{L}}}{\eta_p^2 \theta_1^4 + \theta^2}\right]}$$

In all practical cases $\theta^2 \ll \eta_p^2 \theta_1^4$, so that (1-a) further reduces to the approximate expression

$$\text{(1-b)} \qquad I_1 = \frac{E}{L\omega_o \left[\eta - \frac{\mu K \sqrt{\frac{L_g}{L}}}{\eta_p \theta_1^2} + \frac{\theta^2}{\eta_p \theta_1^4}\right] + \frac{j}{\theta}\left[1 - \frac{\theta^2}{\theta_1^2} - \frac{\mu K \theta^2}{\eta_p^2 \theta_1^4} \sqrt{\frac{L_g}{L}}\right]}$$

This expression gives the variation of I_1 as the circuit is tuned by varying C, the incoming frequency remaining constant, or the variation of I_1 as the incoming wave is varied and the circuit constants remain fixed. In the first case θ_1 is the variable and θ is constant at say the value 1, in which case the incoming wave has wave length λ_o. In the second case, θ is varied while θ_1 remains constant at any chosen value.

Since the arrangement of circuits shown in Figure 1 is seldom used for receiving purposes this equation will not be further investigated.

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CASE 2. OSCILLATORY ELEMENT IN GRID CIRCUIT. In this case the oscillatory element is in the grid circuit, a more usual case when the tube is used as a detector or amplifier.

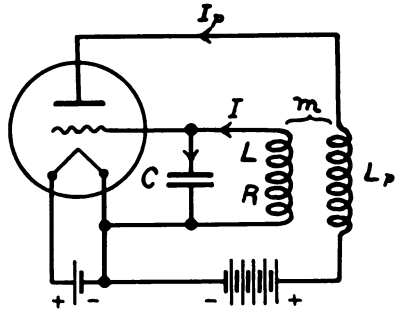

FIGURE 2

Figure 2 gives the connections and the significance of the symbols. The circuit equations for this case are as follows:

$$\begin{cases} \left[R+j\left(L\omega-\dfrac{1}{C\omega}\right)\right]\boldsymbol{I}-jm\omega\boldsymbol{I}_p=\boldsymbol{E} \\ -jm\omega\boldsymbol{I}+(R_p+jL_p\omega)\boldsymbol{I}_p=\mu\boldsymbol{E}_g=-\dfrac{j\mu\boldsymbol{I}}{C\omega} \end{cases}$$

The solution is given in equation (2-1).

$$\boldsymbol{I}=\frac{\boldsymbol{E}}{\left[R-\dfrac{\dfrac{\mu m}{C}-m^2\omega^2}{R_p^2+L_p^2\omega^2}R_p\right]+j\left[\left(L+\dfrac{\dfrac{\mu m}{C}-m^2\omega^2}{R_p^2+L_p^2\omega^2}L_p\right)\omega-\dfrac{1}{C\omega}\right]} \tag{2-1}$$

In this case, since $\dfrac{\mu m}{C}$ is usually greater than $m^2\omega^2$, the effective resistance is reduced and the inductive reactance increased by an increase in m. The circuit oscillates if the effective resistance is made zero or negative. The condition separating oscillation from non-oscillation is

$$R=\frac{\dfrac{\mu m}{C}-m^2\omega^2}{R_p^2+L_p^2\omega^2}R_p.$$

In this case, as in Case 1, the expression for the current given in (2-1) can be expressed in terms of coefficients as in (2-a) below.

$$\boldsymbol{I}=\frac{\boldsymbol{E}}{L\omega_0\left[\gamma-\dfrac{\mu K\theta^2}{\theta_1^2}\dfrac{\sqrt{\dfrac{L}{L_p}}-K^2}{\gamma_p^2\theta^2+1}\gamma_p\right]+\dfrac{j}{\theta}\left[1-\dfrac{\theta^2}{\theta_1^2}+\dfrac{\dfrac{\mu K\theta^2}{\theta_1^2}\sqrt{\dfrac{L}{L_p}}-K^2}{\gamma_p^2\theta^2+1}\right]} \tag{2-a}$$

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If, as is usually the case, $1 \ll \gamma_p^2 \theta^2$, then (2-a) becomes

$$I = \frac{\frac{E}{L\omega_o}}{\left[\gamma - \frac{\mu K}{\gamma_p \theta_1^2}\sqrt{\frac{L}{L_p}} + \frac{K^2}{\gamma_p \theta_1^2}\right] + \frac{j}{\theta}\left[1 - \frac{\theta^2}{\theta_1^2} + \frac{\mu K}{\gamma_p^2 \theta_1^2}\sqrt{\frac{L}{L_p}} - \frac{K^2}{\gamma_p^2 \theta^2}\right]} \quad \text{(2-b)}$$

Referring to equations (1-b) and (2-b) it is evident that the most important resistance-reducing term is not a function of the impressed wave length λ, so that the reduction in resistance for any value of θ_1 is approximately the same for all wave lengths in the emission spectrum of a broadcasting station. This reduction in resistance has roughly the effect of sharpening the resonance curve dependent upon the amount of regeneration, and it is this sharpening of the resonance curve which causes the distortion of signals so familiar to users of the single circuit arrangements when regeneration is pushed to the limit.

If all of the terms of the equations (1-a) and (2-a) be taken into account it is evident that the effective resistance is somewhat a function of the impressed wave length λ. The effective resistance may have slightly negative values for some wave lengths. For these wave lengths the system gives out power when electromotive forces corresponding to these wave lengths are impressed. Under these conditions the system is not an absorber but an emitter for these wave lengths, while for other wave lengths power is absorbed. The field around the receiving antenna may, as a consequence, be distorted differently for different wave lengths so that a neighboring receiving station might receive a distorted signal as a consequence of the activity of the first receiving station. This effect is, however, extremely small for single-circuit arrangements, but is of more importance when coupled circuits are used, as will be pointed out later.

It is interesting to note that the regeneration terms do not in either Case 1 or 2 depend directly upon L, but do depend upon C. In most circuits used for the reception of radio signals, the tuning is accomplished by varying C and the regeneration adjusted with m. The maximum signal is obtained when the reactance term is reduced to zero and the effective resistance reduced to as near zero as possible without oscillation. Both m and C affect both conditions, so that the final adjustment is obtained by alternate adjustments of m and C.

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PART II

THEORY OF COUPLED CIRCUITS

Before discussing the broader question of regeneration in coupled circuits, a brief review will be made of the familiar theory of forced oscillations in two oscillatory circuits having magnetic coupling. Let the two circuits have constants as shown in Figure 3. The impressed electromotive force E has a frequency $\frac{\omega}{2\pi}$. The complex equations (3) below describe completely the conditions of steady state oscillation of this system.

$$(3)\begin{cases} Z_1 I_1 + j M \omega I_2 = E \\ j M \omega I_1 + Z_2 I_2 = 0 \end{cases}$$

where

$$Z_1 = R_1 + j\left(L_1 \omega - \frac{1}{C_1 \omega}\right) = R_1 + j X_1$$

$$Z_2 = R_2 + j\left(L_2 \omega - \frac{1}{C_2 \omega}\right) = R_2 + j X_2$$

Solving these equations, the complex expressions for I_1 and I_2 become

$$I_1 = \frac{E}{Z_1 + \frac{M^2 \omega^2}{Z_2}} = \frac{E}{Z_{12}} = \frac{E}{R_{12} + j X_{12}}$$

(4-1)

$$= \frac{E}{\left[R_1 + \frac{M^2 \omega^2 R^2}{Z_2^2}\right] + j\left[X_1 - \frac{M^2 \omega^2 X_2}{Z_2^2}\right]}$$

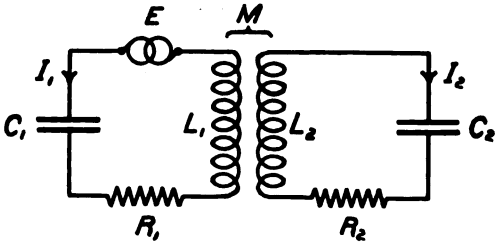

FIGURE 3

where Z_{12} is the impedance of the primary circuit modified by the presence of the coupled secondary circuit, or, in other words, the impedance of the system as viewed from the primary circuit.

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R_{12} and X_{12} are the resistance and reactance of the system as viewed from the primary circuit.

$$(5\text{-}1)\qquad I_2=\frac{-jM\omega E}{Z_1Z_2+M^2\omega^2}=\frac{-jM\omega E}{Z_2\left(Z_1+\dfrac{M^2\omega^2}{Z_2}\right)}=\frac{-jM\omega E}{Z_2Z_{12}}$$

Equations (4-1) and (5-1) can be written in the more general form using coefficients.

$$I_1=\frac{\dfrac{E}{L_1\omega_o}}{\left[\eta_1+\dfrac{\tau^2\eta_2}{\eta^2_2\theta^2+\left(1-\dfrac{\theta^2}{\theta_2{}^2}\right)^2}\right]+\dfrac{j}{\theta}\left[1-\dfrac{\theta^2}{\theta_1{}^2}-\dfrac{\tau^2\left(1-\dfrac{\theta^2}{\theta^2_2}\right)}{\eta_2{}^2\theta^2+\left(1-\dfrac{\theta^2}{\theta^2_2}\right)^2}\right]}\qquad(4\text{-a})$$

$$(5\text{-a})\qquad I_2=\frac{\dfrac{-j\tau\theta E}{\omega_o\sqrt{L_1L_2}}}{\left[\eta_1\theta+j\left(1-\dfrac{\theta^2}{\theta_1{}^2}\right)\right]\left[\eta_2\theta+j\left(1-\dfrac{\theta^2}{\theta_2{}^2}\right)\right]+\tau^2}$$

where $\tau=\dfrac{M}{\sqrt{L_1L_2}}$.

It is convenient to make use of another abbreviation, that is

$$1-\frac{\theta^2}{\theta^2_1}=\beta_1 \text{ and } 1-\frac{\theta_2}{\theta_2{}^2}=\beta_2$$

Using these abbreviations, (4-a) and (5-a) become

$$(4\text{-b})\qquad I_1=\frac{\dfrac{E}{L_1\omega_o}}{\left[\eta_1+\dfrac{\tau^2\eta_2}{\eta^2_2\theta^2+\beta_1{}^2}\right]+\dfrac{j}{\theta}\left[\beta_1-\dfrac{\tau^2\beta_2}{\eta^2_2\theta^2+\beta^2_2}\right]}$$

$$(5\text{-b})\qquad I_2=\frac{\dfrac{-j\tau\theta E}{\omega_o\sqrt{L_1L_2}}}{[\eta_1\theta+j\beta_1]\ [\eta_2\theta+j\beta_2]+\tau^2}$$

Equations (4) and (5) give the primary and secondary currents for any adjustments of the primary and secondary circuits and their coupling. If the coupling between the two circuits is fixed, then X_1 and X_2 are the independent variables and the current can be plotted vertically against the variables X_1 and X_2 (or any other quantities upon which X_1 and X_2 depend), as the two horizontal coordinates. There is thus formed a curved surface bounding a space model. For any value of one of the independent variables there is a *certain* value of the other which will make I_2 or I_1 a maximum or a minimum. If, then, one of the

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variables is given successive values and the other is each time adjusted to give a maximum or a minimum value of the current, there will then result a ridge (if a maximum is obtained) on the representative surface of equation 4 or 5 lying vertically over the corresponding values of X_1 and X_2 which give the maxima. Such curved surfaces with these ridges are shown in Plate 1 for the secondary current I_2.

Since, in most practical cases, the value of the secondary current is of greater interest than that of the primary current, the shape of the representative surface for I_2 will be examined more in detail. The expression (5-1) may be written

$$(6\text{-}1) \qquad I_2 = \frac{-j M \omega E}{Z_1\left(Z_2 + \frac{M^2 \omega^2}{Z_1}\right)} = \frac{-jM\omega E}{Z_1 Z_{21}}$$

$$I_2 = \frac{-j M \omega E}{Z_1\left[\left(R_2 + \frac{M^2 \omega^2 R_1}{R_1^2 + X_1^2}\right) + j\left(X_2 - \frac{M^2 \omega^2 X_1}{R_1^2 + X_1^2}\right)\right]}$$

where Z_{21} is the impedance of the system as viewed from the secondary circuit. Expression (6-1) can be written in terms of coefficients as follows:

$$(6\text{-a}) \qquad I_2 = \frac{\frac{-j\tau\theta E}{\omega_0\sqrt{L_1 L_2}}}{\left[\gamma_1\theta + j\beta_1\right]\left[j\left(\gamma_2\theta + \frac{\tau^2\gamma_1\theta}{\gamma_1^2\theta^2 + \beta_1^2}\right) + j\left(\beta_2 - \frac{\tau^2\beta_1}{\gamma_1^2\theta^2 + \beta_1^2}\right)\right]}$$

It is apparent by inspection of equation (6-1) that if X_1 is set at some value, the value of X_2 which gives a maximum value of I_2 is

$$(7\text{-}1) \qquad X_2 = \frac{M^2\omega^2 X_1}{R_2 + X_1^2} \qquad \text{or}$$

$$(7\text{-a}) \qquad \beta_2 = \frac{\tau^2\beta_1}{\gamma_1\theta + \beta_1^2} \qquad \text{or}$$

$$(7\text{-b}) \qquad 1 - \frac{\theta^2}{\theta_2^2} = \frac{\tau^2\left(1 - \frac{\theta^2}{\theta_1^2}\right)}{\gamma_1^2\theta^2 + \left(1 - \frac{\theta^2}{\theta_1^2}\right)^2}$$

Adjustment of secondary reactance to give max. I_2 max. 1-2 line.

The relation (7) gives a curve on the horizontal or $X_1 - X_2$ plane of the space model above which may be plotted the maximum values of I_2 obtained by first setting X_1 and then adjusting X_2. This locus of max. secondary currents is called max. 1-2 line, and its shape is *independent* of the value of the resistance of the secondary circuit. If, now, this value of X_2 be substituted in

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(6-1), then as X_1 is varied, X_2 is automatically given the correct value to give a maximum I_2, and we follow along the ridge of the space model shown in Plate 1. The maximum value of I_2 is then

$$\text{(8-1)} \qquad max.\ \boldsymbol{I}_2 = \frac{-j\,M\,\omega\,\boldsymbol{E}}{(R_1+jX_1)\left(R_2+\dfrac{M^2\,\omega^2\,R_1}{R_1^2+X_1^2}\right)}$$

or

$$\text{(8-a)} \qquad max.\ \boldsymbol{I}_2 = \frac{\dfrac{-j\,\tau\,\boldsymbol{E}}{\omega_o\sqrt{L_1\,L_2}}}{[M_1\,\theta+j\beta_1]\left[\gamma_2+\dfrac{\tau^2\,\gamma_1}{\gamma_1^2\,\theta+\beta_1^2}\right]}$$

Height of model over max. 1-2 line.

Some value of X_1 will give *the largest* value of $\boldsymbol{I}_2$ attainable (called the max. max. $\boldsymbol{I}_2$ represented by the highest peak of the ridge on the space model. If the complex expression (8) be differentiated to find the value of X_1 which gives this max. max. $\boldsymbol{I}_2$, the following roots result:

$$\text{(9-1)} \qquad X_1 = 0$$

$$\text{(9-2)} \qquad R_1^2+X_1^2 = \frac{M^2\,\omega^2\,R_1}{R_2} \quad \text{or} \quad X_1 = \pm\sqrt{\frac{M^2\,\omega^2\,R_1}{R_2}-R_1^2}$$

conditions for max. max. I_2

$$\text{(9-3)} \qquad R_1^2+X_1^2 = -\frac{M^2\,\omega^2\,R_1}{R_2}$$

(gives imaginary value of X_1)

Root (9-3) gives an imaginary value of X_1. Roots (9-1) and 9-2) give the same value of max. max. $\boldsymbol{I}_2$ if

$$\text{(10-1)} \qquad M^2\,\omega^2 = R_1\,R_2 \quad \text{or}$$

$$\text{(10-a)} \qquad \tau^2 = \gamma_1\,\gamma_2\,\theta^2$$

critical coupling

This latter relation divides the problem into two parts and the coupling which satisfies relation (10) is known as *critical coupling.* For couplings less than critical, root (9-1) gives the max. max. current, the value of which is

$$\text{(11-1)} \qquad max.\ max.\ \boldsymbol{I}_2 = \frac{-j\,M\,\omega\,\boldsymbol{E}}{R_1R_2+M^2\,\omega^2}$$

or

$$\text{(11-a)} \qquad max.\ max.\ \boldsymbol{I}_2 = \frac{-\dfrac{j\,\tau\,\theta\,\boldsymbol{E}}{\omega_o\sqrt{L_1\,L_2}}}{\gamma_1\,\gamma_2\,\theta^2+\tau^2}$$

max. max. I_2 when $\tau <$ critical value

For couplings greater than critical coupling root (9-2) holds. yielding a value for I_2 given by (12)

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$$(12\text{-}1)\qquad max.\ max.\ I_2 = \frac{-j\,M\,\omega\,E}{2\,R_2\left(R_1 \pm j\sqrt{\frac{M^2\,\omega^2\,R_1}{R_2} - R_1^2}\right)}$$

or

$$(12\text{-a})\qquad max.\ max.\ I_2 = \frac{-\,j\,\tau\,\theta\,E}{\omega_o\sqrt{L_1\,L_2}\;2\,\gamma_2\,\theta\left(\gamma_1\,\theta \pm \sqrt{\frac{\tau^2\,\gamma_1}{\gamma_2} - \gamma_1^2\,\theta^2}\right)}$$

max. max. I_2 when $\tau >$ critical value

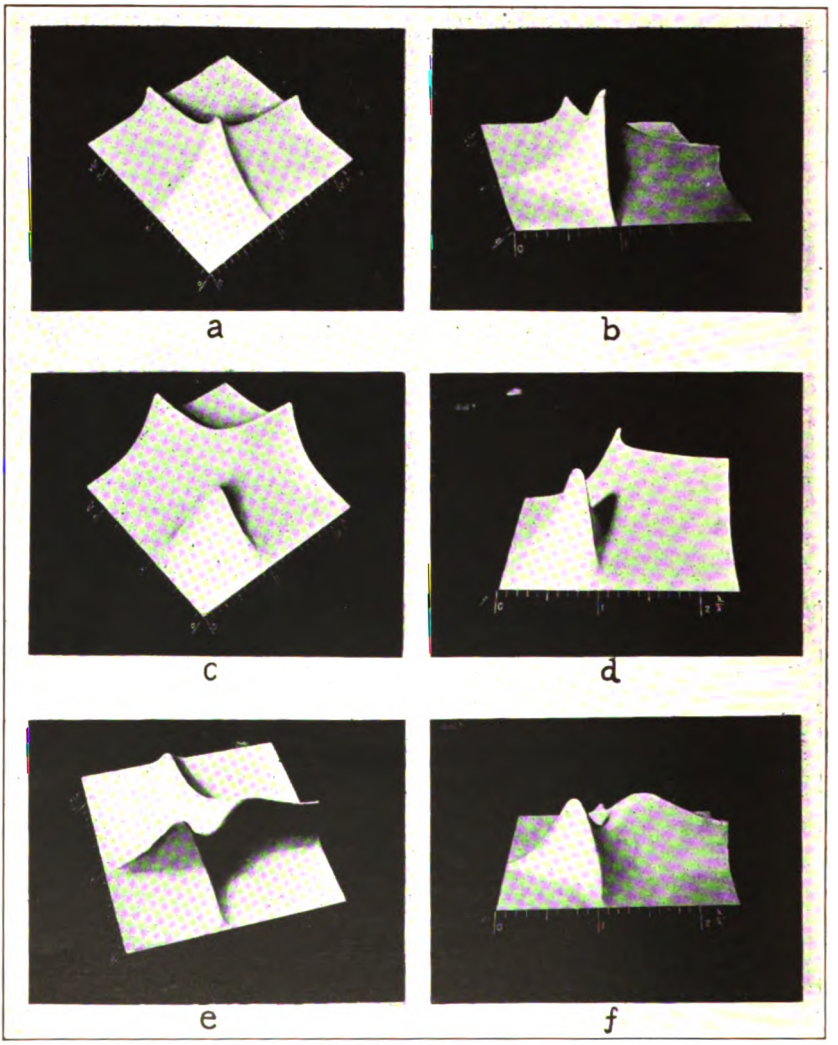

PLATE 1

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This expression for I_2 has a numerical value of

(13-1) $$max.\ max.\ I_2 = \frac{E}{2\sqrt{R_1 R_2}} \quad \text{or}$$

(13-a) $$max.\ max.\ I_2 = \frac{E}{\omega_o \sqrt{L_1 L_2}} \Big/ 2\sqrt{\gamma_1 \gamma_2}$$

max. max. I_2 when $\tau >$ critical value. Numerical value.

In constructing a space model of the secondary current, the ratio of the numerical value of the current to the numerical value of the max. max. current is used. This ratio derived from (5) and (13) is applicable to all cases above critical coupling and including the critical case itself.

(14-1) $$\frac{I_2}{I_{2mm}} = \frac{2 M \omega \sqrt{R_1 R_2}}{\sqrt{(R_1^2+X_1^2)(R_2^2+X_2^2)+M^4\omega^4+ 2M^2\omega^2(R_1 R_2 - X_1 X_2)}}$$

(14-a) $$\frac{I^2}{I_{2mm}} = \frac{2\tau\theta\sqrt{\gamma_1\gamma_2}}{\sqrt{(\gamma_1\gamma_2\theta^2-\beta_1\beta_2+\tau^2)^2+(\gamma_1\theta\beta_2+\gamma_2\theta\beta_1)^2}}$$

$\tau >$ *critical value*

The corresponding expression for this ratio for coupling less than critical is obtained from equation (5) and (11), and is

(15-1) $$\frac{I_2}{I_{2mm}} = \frac{R_1 R_2 + M^2\omega^2}{\sqrt{(R_1^2+X_1^2)(R_2^2+X_2^2)+M^4\omega^4+ 2M^2\omega^2(R_1R_2-X_1X_2)}} \quad \text{or}$$

(15-a) $$\frac{I_2}{I_{2mm}} = \frac{\gamma_1\gamma_2\theta^2+\tau^2}{\sqrt{(\gamma_1\gamma_2\theta^2-\beta_1\beta_2+\tau^2)^2+(\gamma_1\theta\beta_2+\gamma_2\theta\beta_1)^2}}$$

$\tau <$ *critical value.*

In plotting the ratio given in (14) and (15), or any of the other relations, the independent variables β_1 and β_2 may be used instead of the corresponding quantities X_1 and X_2. On the other hand, a better physical picture usually obtains if the quantities $\frac{\omega_1}{\omega}$ and $\frac{\omega_2}{\omega}$, or better $\frac{\theta_1}{\theta} = x$ and $\frac{\theta_2}{\theta} = y$ are used as coordinates.

In studying the action of coupled circuits without regeneration it is usually simpler altho not necessary to consider that the impressed frequency is $\frac{\omega_o}{2\pi}$, making $\theta = 1$.

The surfaces shown in Plate 1 to be described below are equations (14) plotted with x and y as independent variables.

If the coupling between two coupled circuits is less than critical coupling defined by equation (10), then the space model for the

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secondary-current ratio given by equation (15) has a single maximum at values of the independent variables

$$\beta_1 = 0 \quad \text{or } X_1 = 0 \quad \text{or } \frac{\theta_1}{\theta} = \frac{\lambda}{\lambda} = x = 1$$

$$\text{and } \beta_2 = 0 \quad \text{or } X_2 = 0 \quad \text{or } \frac{\theta_2}{\theta} = \frac{\lambda_2}{\lambda} = y = 1$$

The height of this peak is given by equation (15).

At critical coupling the space model has still a single peak, the height of which is now unity. Plate 1, *a* and *b*, shows a calculated model for critical coupling. The peak has a peculiar shape best shown by contour lines given in Figure 4.

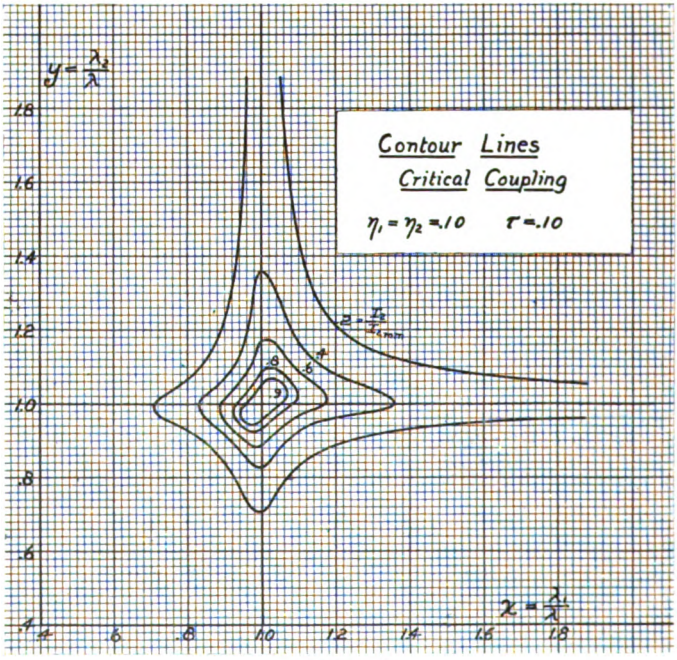

FIGURE 4

If the coupling is greater than critical the space model has in general two ridges, each having a highest point of value unity. Two models calculated from specific values of the coefficients are shown in Plate 1, *c* and *d*, and *e* and *f* (*c* and *d* are two views of the same model and similarly *e* and *f*). As pointed out above, if the secondary circuit is adjusted for a maximum current for successive adjustments of the primary circuit, a locus of points

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on the horizontal plane will be found given by equation (7). Such loci are shown by the curve marked Max. 1-2 (meaning primary set first and then secondary circuit adjusted) in Figure 5 where the coordinates are β_1 and β_2, and in Figure 6 where the same curve is plotted against x and y. It is to be noted that since β_1 is a function of x^2 and β_2 of y^2, all curves plotted to β_1 and β_2 are symmetrical about the origin, but points for β_1 and β_2 greater than $+1$ are imaginary.

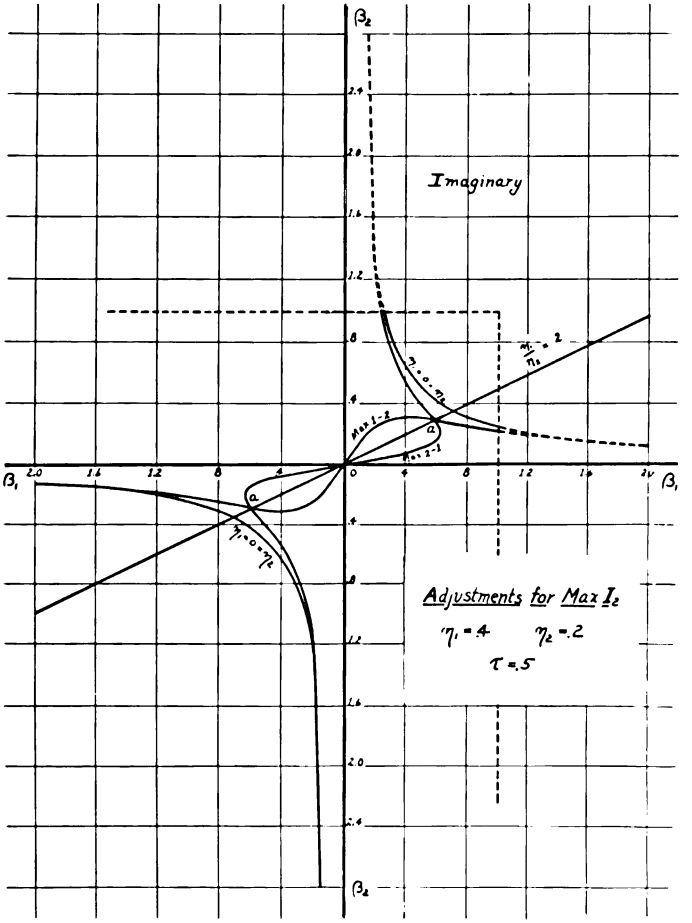

FIGURE 5

It can also be shown that if the reverse order of adjusting the circuits for a maximum is adopted, that is, if the primary circuit is adjusted for a maximum secondary current for every adjust-

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ment of the secondary circuit, a new curve will be obtained given by the equation.

(16-1) $$X_1 = \frac{M^2 \omega^2 X_2}{R_2^2 + X_2^2}$$ or

(16-a) $$\beta_1 = \frac{\tau^2 \beta_2}{\eta^2_2 + \beta_2^2}$$

Adjustment of primary circuit for max. I_2. Max. 2-1 line

This locus is shown by the curve Max. 2-1 in Figure 5 and Figure 6. The intersections of Max. 1-2 and Max. 2-1 lines shown at *a a*, Figure 6, gives the positions of the max. max. current, the coordinates of which are given by equation 9. (If $\eta_1 = 0 = \eta_2$ curves Max. 1-2 and Max. 2-1 resolve into the curves so marked in the figures. At critical coupling the two intersection points

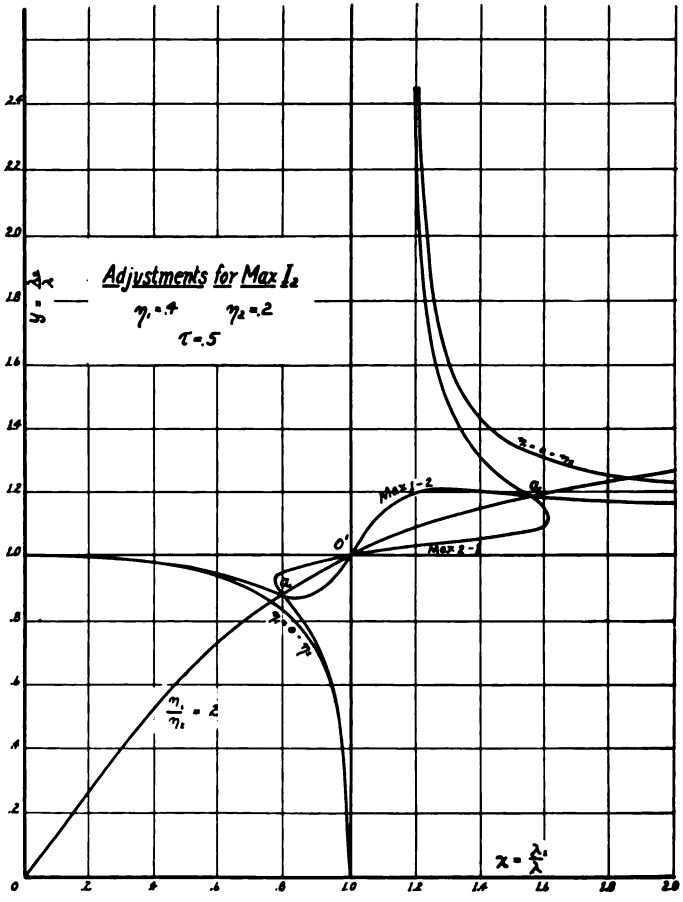

FIGURE 6

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a a of Figures 6 and 7 merge into a single point at the origin as shown in Figure 7. If $\eta_1 = \eta_2$, then these curves Max. 1-2 and Max. 2-1 will be similar in shape and symmetrical about the 45° line, but otherwise the two sets will be dissimilar as shown in the figures.

The reason the two sets of curves are different according to the order of adjustment will be apparent if the shape of the saddle between the two peaks (Plate 1, *c*, *d*, *e*, and *f*) is considered. A card with its plane vertical and its lower edge horizontal and parallel to the y axis will rest on the surface at one point. If the card is now moved in the x direction this point of tangency will travel over the Max. 1-2 line. The Max. 2-1 line can be traced by the card placed at right angles to the y axis and moved in the direction of the y axis.

For any particular coupling the highest peaks have definite coordinates. If, now, the coupling be changed, the two peaks will move along a curve gotten by eliminating the coupling from the two equations (7) and (9-2), giving

$$\frac{R_1}{R_2} = \frac{X_1}{X_2} \quad \text{or} \quad \frac{\eta_1}{\eta_2} = \frac{\beta_1}{\beta_2} \qquad \left.\begin{array}{l}\\ \\ \end{array}\right\} \textit{locus of peaks as coupling is changed}$$

(17-1)

(17-a)

This equation is plotted in Figure 6 for the particular value $\frac{\eta_1}{\eta_2} = 2$ and for other values of η_1 in Figure 8. The dotted line in Figure 8 gives the Max. I_2 line for $\tau = \frac{\eta_1}{\eta_2}$ and is hence the limit of the peak loci.

It can be shown that for all values of coupling the ridge along the saddle between the two peaks is directly above the curve of (17) for the appropriate value of $\frac{\eta_1}{\eta_2}$.

Attention is now called to the fact that if η_1 is equal to η_2 the model is symmetrical with respect to the coordinate axes and the peaks lie over the 45° line, that is, over the line $\beta_1 = \beta_2$ or $x = y$. (See Plate 1, *c* and *d*.) If on the other hand η_1 is *not* equal to η_2 then the peaks are skewed around as shown in Plate 1, *e* and *f*. In practice the primary circuit may be the antenna circuit and the secondary circuit of the present discussion the secondary circuit of a coupled-circuit receiver. In almost all practical cases the decrement coefficient of the antenna circuit, due to the ground and radiation resistance, is many times that of the secondary circuit. Especially is this true when regeneration is

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used in the secondary circuit. It is evident then that the space model of the practical receiver is very much distorted, even more so than represented by the model in Plate 1, *e* and *f*.

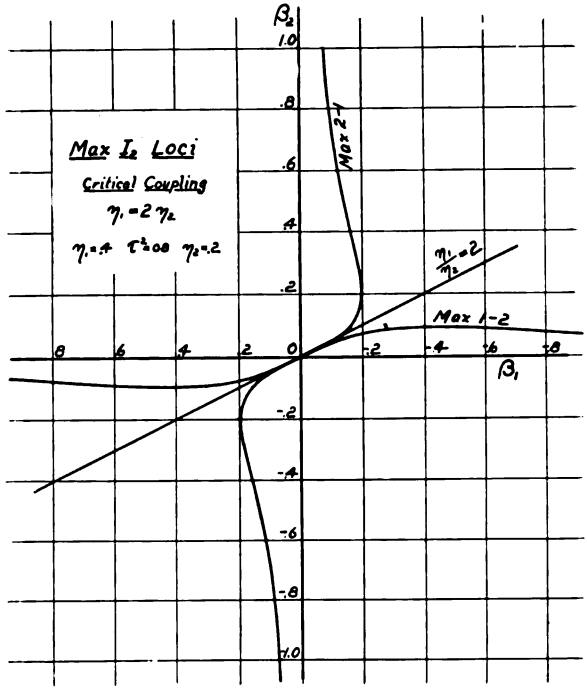

FIGURE 7

It is well now to examine, by means of these space models, the selectivity and tuning relations of a coupled-circuit receiver. If the circuits are adjusted to any point on the x y-plane and then the natural wave length of say, the primary circuit, is varied while a signal of constant wave length $y = y_o$ is coming in, the variation in secondary current is given by the cross section of the model along a line thru the original point parallel to the x axis. Similarly a cross section parallel to the y axis gives the variation of current when the secondary circuit reactance is varied. If, on the other hand, the two circuits are fixed and the incoming wave length y is changed, the resulting variation of secondary current can be obtained from equation (14-a) or (15-a) by varying θ, remembering that β_1 and β_2 are both functions of θ. If θ occurred nowhere in these equations except in β_1 and β_2, then since θ appears in β_1 and β_2 symmetrically, the variation in I_2

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would be given by allowing x and y to vary proportionately, that is by taking a cross section thru the space model along a radial line thru any point determined by the settings of θ_1 and θ_2. The fact that θ does occur in other terms of (14-a) and (15-a) makes this method of obtaining the result incorrect. Nevertheless, the terms other than β_1 and β_2 which contain θ are slowly varying terms compared to β_1 and β_2 so that the cross section obtained as outlined above does, after all, give a result near enough to the correct result to make the method a very useful one. The smaller γ_1 and γ_2 are, and the less the variation in θ necessary to obtain the required resonance curves, the truer is the result obtained by this cross-sectioning method.

In using this method it must be remembered, however, that since the variation in the coordinated x and y in this case is *inversely* proportional to y, the shape of the curve is somewhat distorted from what it would be if plotted in the usual way against y directly, moving along the radial line toward the origin means an *increase* in y. If the distance moved is very small the distortion due to this effect is small.

Examining now a specific case, let $\gamma_1 = 5\ \gamma_2$. The factor of proportionality 5 is much less than is often met in practice. Because

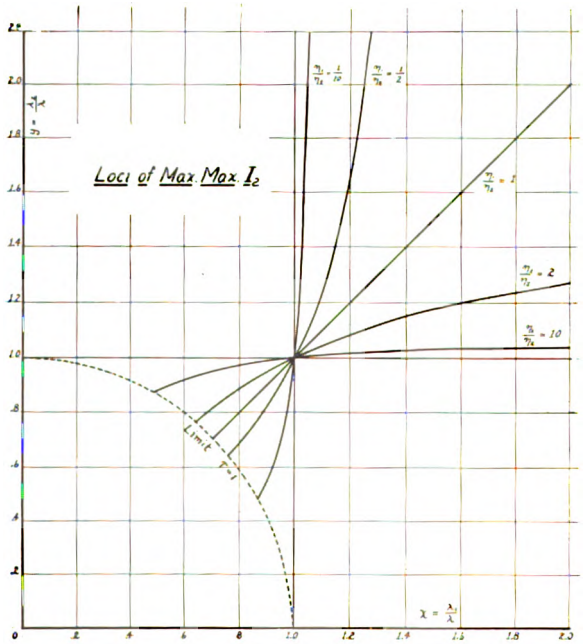

FIGURE 8

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of distributed capacity of the coils a stronger signal is often obtained when the coupling is above critical and the adjustments are made for the long wave length maximum. This gives sufficient reason for considering the case shown in Plate 1, *e* and *f*, a practical case. It is evident that the primary circuit is adjusted to a much longer wave length than the resonant value for the incoming wave and that the apparent tuning in the primary circuit is very dull as shown by the section *pp* parallel to the *x* axis of the space model shown in Figure 9. The tuning in the secondary circuit is sharp as indicated by the section *ss*. The selectivity against interference from other stations, is, however, relatively low as indicated by the diagonal section along the line *oo*. The corresponding sections *p'p'* and *s's'* through the other peak of the space model are also shown in Figure 9.

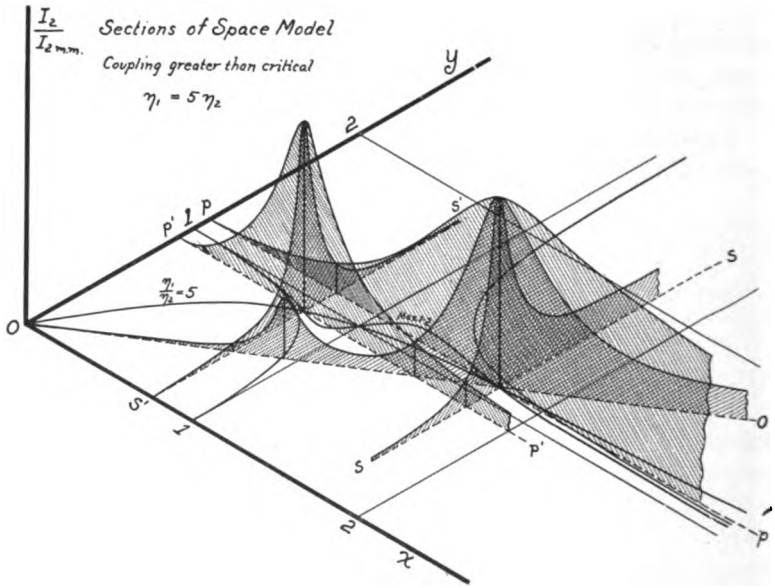

FIGURE 9

Figure 10 is a sectional drawing of the special case when η_1 is equal to η_2. As pointed out above, the space model under these conditions is symmetrical as shown in *c* and *d* of Plate 1 and the two max. max. current peaks lie over the 45° line of the *x*-*y* plane.

The above discussion indicates the manner in which the space model can be used to advantage to study the equivalent resonance curve, and the selectivity and tuning relations under different conditions of the citcuits. Enough has been given to show

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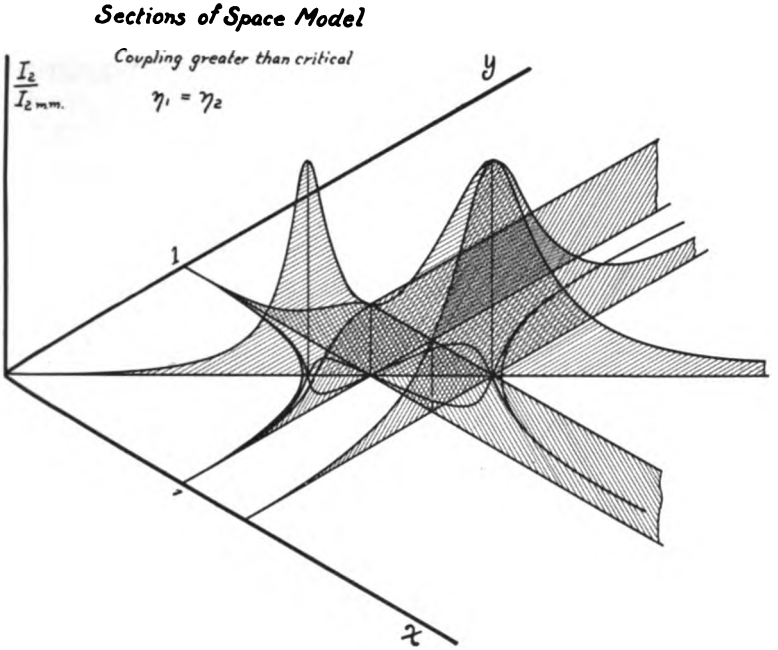

FIGURE 10

that the ratio of values of the decrement coefficients as well as their absolute values is important.

Referring now to the case of critical coupling, there is only one main maximum, but if the circuits are not in resonance, there are two values of the incoming wave length which will give maxima. This is shown by the section *oo'* and *oo''* in Figure 11, which shows cross sections of the space model for critical coupling when $\gamma_1 > \bar{\gamma}_2$. Again, it is interesting to note that the equivalent resonance curve at critical coupling when the incoming wave is varied, is broad at the top as shown in Figure 12, whereas the curve obtained for a fixed y thru the point $x = 1$ and $y = 1$ as either y_1 or y_2 is varied, is sharper. (See Figure 4.) This broadness of resonance with a flat top is of advantage when receiving a radio telephone having a long carrier wave because of the broad spectrum which must be received.

The reason for the broader shape of the critical coupling resonance curve as y is varied is easily seen when it is remembered that for all couplings above zero, coupled circuits have two free periods of oscillation and there are two branches to the equi-coupling curves similar to that shown in Figure 6. A cross section along the 45° line thru these branches obtained by varying y gives two

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peaks of current. For couplings above critical the two current peaks can be resolved, but at and below critical coupling the peaks are so close that they merge and cannot be resolved. The resultant resonance curve is, therefore, a sum of two curves very close together and hence is, necessarily, broader than a single resonance curve.

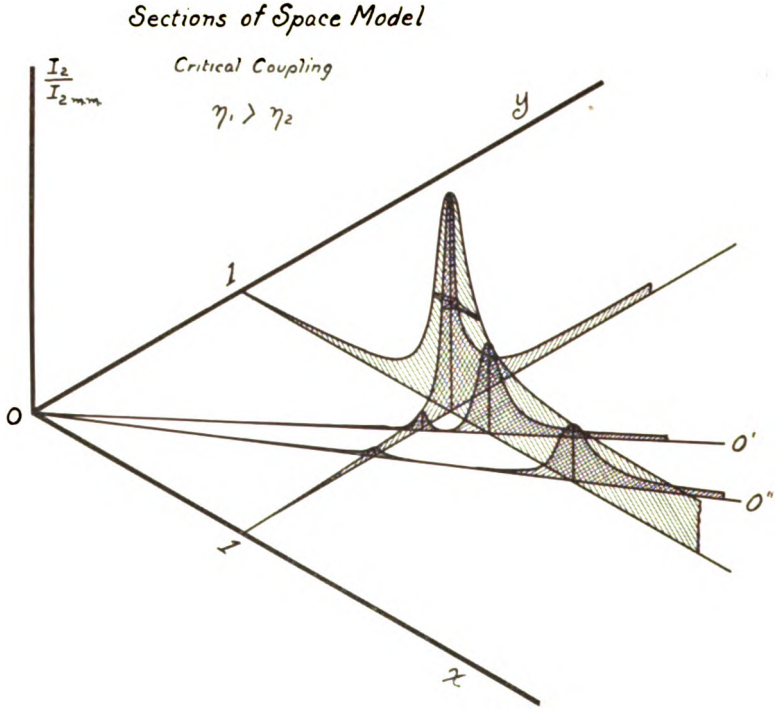

FIGURE 11

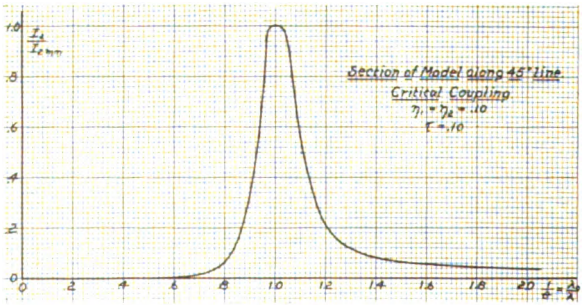

FIGURE 12

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In the further development of radio communication it is conceivable that it will be desirable to receive more complex radio spectra such, for instance, as two waves with considerable separation. Such a spectrum can be received by making the two peaks of coupled circuits coincide with the two waves, but this can be done efficiently only when the two maxima occur on the 45° line of the model, as shown in Figure 10, and this is possible only when the ratio of resistance to inductance for the two circuits is the same. Since, as has been pointed out, the resistance of the antenna circuit is usually much greater than that of the secondary circuit, the reception of a two-wave spectrum necessitates the reduction of the primary circuit resistance by some method such as regeneration in that circuit.

Part III

Regeneration in Coupled Circuits

The familiar theory of coupled circuits presented above has been given as a necessary foundation to the following theory of regeneration in coupled circuits. The author apologizes for the seemingly necessary repetition, but it is hoped that possibly the method of presentation may be of interest.

The theory of regeneration in coupled circuits is necessarily complex and a complete solution will not be attempted. Any one who has operated a double circuit regenerative receiver knows the complexity of adjustments and the interdependence of every adjustment on every other, so it is believed that a discussion of the theory of the case may enlighten the operations.

The only case considered in this paper is that in which regeneration takes place in one of two coupled circuits having constants shown in Figure 13. The circuit having regeneration is denoted by subscripts 2 and is termed the secondary circuit altho this terminology is entirely arbitrary. The emf. may be induced in either circuit. The equations for this case are readily deduced and are as follows:

$$(18) \qquad \begin{cases} \boldsymbol{Z}_1 \boldsymbol{I}_1 + j M \omega \boldsymbol{I}_2 = \boldsymbol{E}_1 \\ j M \omega \boldsymbol{I}_1 + \boldsymbol{Z}_2 \boldsymbol{I}_2 - j M \omega \boldsymbol{I}_p = \boldsymbol{E}_2 \\ j \left(\frac{\mu}{C_2 \omega} - m \omega \right) \boldsymbol{I}_2 + \boldsymbol{Z}_p \boldsymbol{I}_p = 0 \end{cases}$$

where

$$\boldsymbol{Z}_1 = R_1 + j \left(L_1 \omega - \frac{1}{C_1 \omega} \right)$$

$$\boldsymbol{Z}_2 = R_2 + j \left(L_2 \omega - \frac{1}{C_2 \omega} \right)$$

$$\boldsymbol{Z}_p = R_p + j L_p \omega$$

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M is the mutual inductance between L_1 and L_2, m the mutual inductance between L_2 and L_p, and μ is the amplification factor of the tube. E_1 and E_2 are emfs. of frequency $\frac{\omega}{2\pi}$ introduced into the first and second circuits, respectively. In most of the following discussion it is simpler to consider the impressed frequency of the emf. E_1 and E_2 to be equal to the reference frequency $\frac{\omega_o}{2\pi}$, except in those discussions where the impressed frequency is assumed to vary for the purpose of obtaining the shape of the resonance curve of the system. The equations will be developed, however, without this limitation and they are expressed in terms of any frequency $\frac{\omega}{2\pi}$.

The solution of equation (18) is given below

$$I_1 = \frac{E_1\left(Z_2 - \frac{\frac{\mu m}{C_2} - m^2\omega^2}{Z_v}\right) - jM\omega E_2}{Z_1\left(Z_2 - \frac{\frac{\mu m}{C_2} - m^2\omega^2}{Z_p}\right) + M^2\omega^2} \tag{19}$$

$$I_2 = \frac{E_2 Z_1 - jM\omega E_1}{Z_1\left(Z_2 - \frac{\frac{\mu m}{C_2} - m^2\omega^2}{Z_p}\right) + M^2\omega^2} \tag{20}$$

It is apparent on examination of equations (19) and (20) that the coupling m effects only Z_2, and hence regeneration in a circuit affects *only* the impedance of that circuit and not the impedance of any circuit coupled to it.[1] The equivalent impedance of the second circuit or the circuit in which the regeneration is established is

$$\bar{Z}_2 = Z_2 - \frac{\frac{\mu m}{C_2} - m^2\omega^2}{Z_p} \quad \text{or} \tag{21}$$

Equivalent impedance of regenerated circuit

$$\bar{Z}_2 = \left[R_2 - \frac{\frac{\mu m}{C_2} - m^2\omega^2}{Z_p^2} R_p\right] + j\left[X_2 + \frac{\frac{\mu m}{C_2} - m^2\omega^2}{Z_p^2} X_p\right] \quad \text{or} \tag{22}$$

$$\bar{Z}_2 = [R_2 - H R_p] + j[X_2 + H X_p] \quad \text{or} \tag{23}$$

$$\bar{Z}_2 = \bar{R}_2 + j\bar{X}_2 \tag{24}$$

[1] Several statements have appeared in the literature which claim that regeneration in the secondary circuit of two coupled circuits reduces the resistance of the primary or antenna circuit. Such statements are incorrect.

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In equation 23 the quantity $\frac{\frac{\mu m}{C_2} - m^2\omega^2}{Z_p^2}$ has been replaced by the abbreviation H, which will be called the *coefficient of regeneration.* H has no dimensions and is nearly independent of frequency, particularly if C_2 and L_p are small. Expressed in terms of coefficients, H has the value

(25-1) $$H = \frac{\frac{\mu m}{C_2} - m^2\omega^2}{Z_p^2} \quad \text{or}$$

(25-a) $$H = \frac{L_2}{L_p} \cdot \frac{k\left(\frac{\mu}{y^2}\sqrt{\frac{L_2}{L_p}} - k\right)}{\gamma_p^2\theta^2 + 1}$$

coefficient of regeneration

where $y = \frac{\frac{\lambda_2}{\lambda_0}}{\frac{\lambda}{\lambda_0}}$ as before, and $k = \frac{m}{\sqrt{L_p L_2}}$ is the coefficient of coupling between L_p and L_2.

Examination of equation (25) shows that regeneration gives H a positive value unless $m^2\omega^2 > \frac{\mu m}{C_2}$, which is seldom the case. Since H depends upon m, the effect of m or k is to alter the values of resistance and reactance of the *regenerated circuit only.* The resistance R_2 is changed to an effective or equivalent value $\bar{R}_2$ and the reactance X_2 to an effective value $\bar{X}_2$. The change in the reactance can be attributed to an apparent change in L_2 as developed in the following equations.

(26-1) $$\bar{X}_2 = X_2 + H X_p$$

(26-2) $$= (L_2 + H L_p)\,\omega - \frac{1}{C_2\,\omega}$$

(26-3) $$= L_2\,\omega - \frac{1}{C_2\,\omega} \qquad \text{where } \bar{L}_2 \text{ is the equivalent inductance.}$$

Regeneration increases the apparent inductance to $L_2 + H L_p$. In practice the actual change is small.

The following quantities may now be defined in terms of $\bar{L}_2$.

(27) $$\bar{\omega}_2 = \frac{1}{\sqrt{\bar{L}_2 C_2}}$$

(28) $$\bar{\lambda}_2 = \frac{2\pi V}{\bar{\omega}_2} \qquad V = \text{velocity of light.}$$

(29) $$\bar{\beta}_2 = 1 - \frac{1}{\left(\frac{\bar{\lambda}_2}{\lambda}\right)^2} = 1 - \frac{\theta^2}{\bar{\theta}_2^2}$$

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(30) $$\bar{\tau} = \frac{M}{\sqrt{L_1 \bar{L}_2}}$$

(31) $$\bar{K} = \frac{m}{\sqrt{L_p \bar{L}_2}}$$

Then it follows that since

(32) $$\bar{R}_2 = R_2 - H R_p, \text{ then}$$

(33) $$\bar{\gamma}_2 = \frac{\bar{R}_2}{\bar{L}_2 \omega_o} = \gamma_2 \frac{L_2}{\bar{L}_2} - H \frac{L_p}{\bar{L}_2} \gamma_p$$

The expression for H can be given in terms of equivalent values

(25-b) $$H = \frac{\bar{L}_2}{L_p} \cdot \frac{\bar{K}\left(\frac{\mu}{\gamma_2}\sqrt{\frac{\bar{L}_2}{Q_p}} - \bar{K}\right)}{\gamma_p^2 \theta^2 + 1}$$

In discussing the conditions of oscillation and maximum regeneration it is easiest to assume first that E_2 is zero. This case is the most practical, for E_1 may then be considered to be the signal emf. induced in the antenna circuit represented by the primary circuit of Figure 13, and the secondary circuit of the same figure is the usual closed secondary circuit with regenerative detector or amplifier. Equations (19) and (20) then reduce to

(34) $$I_1 = \frac{E_1 \bar{Z}_2}{Z_1 \bar{Z}_2 + M^2 \omega^2} = \frac{E_1 [(R_2 - H R_p) + j(X_2 + H X_p)]}{Z_1 [(R_2 - H R_p) + j(X_2 + H X_p)] + M^2 \omega^2}$$
$$= \frac{E_1}{Z_1 + \frac{M^2 \omega^2}{\bar{Z}_2}} = \frac{E_1}{\bar{Z}_{12}}$$

where $\bar{Z}_{12}$ is the equivalent impedance of the system looking from the primary circuit.

(35)
$$I_2 = -\frac{j M \omega E_1}{Z_1 \bar{Z}_2 + M^2 \omega^2} = -\frac{j M \omega E_1}{Z_1 [(R_2 - H R_p) + j(X_2 + H X_p)] + M^2 \omega^2}$$
$$= \frac{-j M \omega E_1}{\bar{Z}_2 \left(Z_1 + \frac{M^2 \omega^2}{\bar{Z}_2}\right)} = \frac{-j M \omega E_1}{Z_2 Z_{12}} = \frac{-j M \omega E_1}{Z_1 \left(\bar{Z}_2 + \frac{M^2 \omega^2}{Z_1}\right)} = \frac{-j M \omega E_1}{Z_1 \bar{Z}_{21}}$$

where $\bar{Z}_{21}$ is the equivalent impedance of the system looking from the secondary circuit.

If now $H R_p$ is made equal to or greater than R_2, then R_2 will become zero or negative. If the regenerated circuit is not coupled to the first circuit (that is if $M = 0$) the second circuit will oscillate as soon as R_2 becomes zero or negative. This condition that $R_2 = 0$ is the same as the condition of oscillation of the simple

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circuit of Case 2, Part I. When, however, the two circuits are coupled, $\bar{R}_2$ may be negative without oscillations taking place, because of the additional losses introduced by the primary circuit. There is then presented the peculiar condition of two coupled circuits, the effective resistance of one of which may be negative. It was this interesting condition which partially induced the present investigation in order to determine how the usual relations of coupled circuits with positive resistances given in the second part of this paper would be altered.

Case 1. Special Care of Constant Regenerative Effect. Referring to equation (25) it is evident that the regeneration coefficient H is a function of both θ_2 and θ, that is, H varies with the setting of the secondary condenser and the incoming wave length, altho the variation with the latter is slight. The first case to be considered assumes H is constant. Under this assumption the resistance of the secondary circuit is altered to a constant value $\bar{R}_2$, which may be negative and L_2 is slightly altered to a new equivalent value $\bar{L}_2$. All quantities will be expressed and plotted in terms of the equivalent values. This case applies to two circuits, one having a constant negative resistance obtained in any way, and also to the practical case of coupled circuits having a regenerative secondary circuit when the tuning is done by varying L_2 so that m does not change.

To investigate further, the same procedure used in the theory of coupled circuits with positive resistances may be followed. The secondary current given by equation (35) is expanded below in terms of equivalent values and is similar to equation (6) for coupled circuits with positive resistances.

$$\text{(36-1)} \qquad \boldsymbol{I}_2 = \frac{-j\,M\,\omega\,\boldsymbol{E}_1}{\boldsymbol{Z}_1\left[\left(R_2 + \frac{M^2\,\omega^2\,R_1}{R_1^2+X_1^2}\right)+j\left(X_2-\frac{M^2\,\omega^2\,X_1}{R_1^2+X_1^2}\right)\right]} \quad \text{or}$$

$$\text{(36-a)} \qquad \boldsymbol{I}_2 = \frac{-j\,\tau\,\theta\,\boldsymbol{E}_1\,\omega_0\sqrt{L_1\,L_2}}{[\gamma_1\theta+j\,\beta_1]\left[\left(\gamma_2\theta+\frac{\tau^2\,\gamma_1\,\theta}{\gamma_1^2\,\theta^2+\beta_1^2}\right)+j\left(\beta_2-\frac{\tau^2\,\beta_1}{\gamma_1^2\theta^2+\beta_1^2}\right)\right]}$$

Now referring to equation (36-1) and following the same procedure as before, if X_1 has some value and X_2 is varied until a maximum of $\boldsymbol{I}_2$ is obtained, the value of X_2 is

$$\text{(37-1)} \qquad X_2 = \frac{M^2\,\omega^2\,X_1}{R_1^2+X_1^2} \quad \text{or} \quad \textit{X_2 to give max. I_2 equation of max. 1-2 line}$$

$$\text{(37-a)} \qquad \beta_2 = \frac{\tau^2\,\beta_1}{\gamma_1^2\,\theta^2+\beta_1^2} \qquad \textit{β_2 to give max. I_2, equation of max. 1-2 line}$$

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Equation (37) corresponds to equation (7) and is *independent* of the value of R_2, whether negative or positive.

The value of the max. I_2 is

$$\text{(38-1)} \qquad max.\ I_2 = \frac{-j\,M\,\omega\,E_1}{(R_1+jX_1)\left(\bar{R}_2+\dfrac{M^2\omega^2 R_1}{R_1^2+X_1^2}\right)} \quad \text{or}$$

$$\text{(38-a)} \qquad max.\ I_2 = \frac{-\dfrac{j\bar{\tau}\theta E_1}{\omega_0\sqrt{L_1\bar{L}_2}}}{[\gamma_1\theta+j\beta_1]\left[\bar{\gamma}_2\theta+\dfrac{\bar{\tau}^2\gamma_1\theta}{\gamma_1^2\theta^2+\beta_1^2}\right]}$$

Obtained by adjusting X_1 then $\bar{X}_2$

This expression evidently gives a finite value of I_2 even if $\bar{R}_2$ is negative, but less in absolute value than $\dfrac{M^2\omega^2 R_1}{R_1^2+X_1^2}$. The conditions under which I_2 shall have its greatest value can now be found, as was done in the simple theory of coupled circuits, by differentiating equation (38-1), remembering it is a complex expression. The roots thus found for X_1 are similar to equation (9) in Part II.

$$\text{(39-1)} \qquad X_1 = 0$$

$$\text{(39-2)} \qquad R_1^2+X_1^2 = \frac{M^2\omega^2 R_1}{\bar{R}_2} \qquad \textit{Imaginary value of } X_1 \textit{ if } \bar{R}_2 \textit{ is negative.}$$

$$\text{(39-3)} \qquad R_1^2+X_1^2 = -\frac{M^2\omega^2 R_1}{\bar{R}_2} \qquad \textit{Value of } X_1 \textit{ to give max. max. } I_2.$$

Expression (39-2), which in Part II gave the value of X_1 for the max. max I_2, now, when $\bar{R}_2$ is negative, gives an imaginary value. Therefore root (39-3) must now be used.

Examining more closely the conditions given in (39) it is evident that (39-3) gives in general for any value of ω two values of X_1 for which the denominator of equation (38) is *zero*. At these two values of X_1 the current I_2 (and also I_1) becomes infinite or oscillation begins at frequency $\dfrac{\omega}{2\pi}$. Values of X_1 (with corresponding values of X_2, since the discussion assumes that adjustments are such that equation (37) holds) between these two points give finite values of I_2 of frequency $\dfrac{\omega}{2\pi}$. Since the only adjustments under considerations lie on the Max. 1-2 line shown in Figures 5 and 6 for $\theta = 1$, this line included between the boundary points of oscillation at frequency $\dfrac{\omega}{2\pi}$ $\left(\dfrac{\omega_o}{2\pi}\right.$ for Figures 5 and 6,

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since θ equals unity for these figures), is the locus of max. I_2 just as in the second part of this paper when the secondary resistance was positive. It should be noted that the shape of the Max. 1-2 line is independent of whether $\overline{R}_2$ is positive or negative and that the *same* two values of X_1 satisfy equation (9-2) that satisfy (39-3) if the negative $\overline{R}_2$ has the same numerical value that the positive R_2 had in Part II.

As was demonstrated in Part II, the Max. 2-1 line (adjustment of primary circuit to max I_2 for various settings of secondary circuit) passes thru the points of max. max. I_2, which have in this case of negative resistance been shown to be the boundary points between oscillation and non-oscillation. The portion of this Max. 2-1 line between these two points is in this case the locus of max. I_2 of frequency $\frac{\omega}{2\pi}\left(\frac{\omega_o}{2\pi}\text{ if }\theta=1\right)$ where the order of adjustment of the two circuits is as indicated, that is, first secondary then primary.

For points on the Max. 1-2 line outside the region between the boundary points of oscillation, $\overline{R}_2+\frac{M^2\omega^2 R_1}{R_1^2+X_1^2}$ (see equation (36)) is negative and $\overline{X}_2-\frac{M^2\omega^2 X_1}{R_1^2+X_1^2}=0$. This condition cannot exist because if the first expressions were negative, a larger amount of power would be supplied to the system than would be dissipated, an obvious impossibility. What actually happens is first an increase in magnitude of the oscillation. The vacuum tube then operates over such an extent of its characteristics that the average values of μ and R_p are altered. This alteration in μ and R_p is, in each case, in the direction of reducing the negativity of $\overline{R}_2+\frac{M^2\omega^2 R_1}{R_1^2+X_1^2}$. Equilibrium is established when this expression again becomes zero. Of course harmonics are introduced which complicate matters, but to a first approximation in the qualitative discussion they may be neglected. The variations in μ and R_p will also alter the value of $\overline{X}_2$, but to the first approximation, equation (37) may still be considered true and the oscillation at frequency $\frac{\omega}{2\pi}\left(\frac{\omega_o}{2\pi}\text{ if }\theta=1\text{ as in Figures 5 and 6}\right)$ will then take place over the portions of the Max. 1-2 line outside the boundary points.

If equations (37) and (39-3), both of which hold for these two points of boundary between oscillation and non-oscillations, are solved simultaneously there results

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(40-1) $\dfrac{R_1}{\bar{R}_2} = -\dfrac{X_1}{\bar{X}_2}$ or

(40-a) $\dfrac{\eta_1}{\bar{\eta}_2} = -\dfrac{\beta_1}{\bar{\beta}_2}$

Locus of points where oscillation begins at frequency $\dfrac{\omega_o}{2\pi}$ *as coupling M is varied.*

Equation (40) is evidently the same as (17) in Part II which is the equation of the straight lines in Figure 5 and Figure 7, and the curved lines in Figure 6 and Figure 8. Attention is again called to the fact that the figures are all drawn for the case that $\theta = 1$. The boundary points given by equation (39-3) are therefore, as in Part II, found to be the points of intersection of the Max. 1-2 line, the Max. 2-1 line and the appropriate radial line. These points of intersection which now mark the boundary between oscillation and non-oscillation were in Part II the points of max. max. I_2, the current in the present case increasing toward infinity instead of to a finite max. max. value.

If, now,

(41-1) $M^2 \omega^2 = -R_1 \bar{R}_2$ or

(41-a) $\tau^2 = -\eta_1 \bar{\eta}_2 \theta^2$

Critical coupling.

a relation which gave critical coupling in Part II, the boundary points just discussed come together at $X_1 = 0$ and $\bar{X}_2 = 0$. Relation (41) marks then a sort of critical coupling for negative resistance in one circuit, for then the region of no oscillation at frequency $\dfrac{\omega}{2\pi}$ has shrunk to a point and become a point of max. max. I_2, that is, infinite I_2. This adjustment should be the condition for maximum signal with greatest selectivity. A little analysis shows that under this condition the line Max. 1-2 and the locus of equation (40) are tangent at the origin as shown in Figure 7, which is drawn for the case $\theta = 1$.

It is, therefore, evident from the above discussion that when two circuits are coupled and one, say the secondary, has regeneration so that its effective resistance is negative but less in absolute value than $\dfrac{M^2 \omega^2 R_1}{R^2_1 + X_1^2}$, and if the new equivalent reactance determines the new natural wave length $\bar{\lambda}_2$ which will be slightly different from λ_2, and if now all equations be plotted to $x = \dfrac{\eta_1}{\theta}$ and $y = \dfrac{\bar{\theta}_2}{\theta}$ or to β_1 and $\bar{\beta}_2$, then the locus Max. 1-2 (shown for $\theta = 1$ in Figure 5), which marks the maximum value of I_2 in simple coupled circuits when X_1 and then X_2 are adjusted, also marks the maximum value of current of frequency $\dfrac{\omega}{2\pi}$ when the second-

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ary resistance is negative. As this locus curve is traversed, there are two values of X_1 marking a range outside of which the system oscillates at frequency $\frac{\omega}{2\pi}$. Between the two values of X_1 the points on curve Max. 1-2 give max. I_2 unless the tube starts oscillating at some other frequency, a possibility which will be shortly explained.

Referring to Figure 6, for which $\theta = 1$ and which may now be used for the case of negative resistance $\bar{R}_2$, oscillation at frequency $\frac{\omega_o}{2\pi}$ takes place if the adjustments are such as to correspond to points on the locus Max. 1-2 outside the region between the intersections of the Max. 1-2 and the Max. 2-1 lines. Over the portions of the Max. 1-2 and the Max. 2-1 lines between these points of intersection, the current I_2 will be a maximum when the adjustments are made in the proper order. It is evident that maximum regeneration at frequency $\frac{\omega_o}{2\pi}$ will occur at the points of intersection of the Max. 1-2 and Max. 2-1 lines, that is, at points *a-a*.

It is apparent that if M is varied, the points of intersection (points of max. max. I_2) travel along the $\frac{\eta_1}{\eta_2}$ line. If, however, M is fixed and $\bar{R}_2$ varied, the points *a-a* travel along the Max. 1-2 line which is itself unchanged in shape. If $\bar{R}_2$ is positive, the points *a-a* represent points of max. max. I_2. As $\bar{R}_2$ approaches zero the points recede from the origin 0′ along the Max. 1-2 line until when $\bar{R}_2 = 0$ the points are at $x = 0$ and $x = \infty$. As $\bar{R}_2$ becomes a larger negative quantity, the points *a-a* approach 0′ along the same Max. 1-2 line, this time marking points where oscillation at frequency $\frac{\omega_o}{2\pi}$ begins, or points of maximum regeneration. At these points I_2 is theoretically infinite.

The boundary line on the *x-y* plane separating the region of non-oscillation from that where oscillation at *some* frequency takes place can now be deduced. It must be remembered that whereas frequency $\frac{\omega}{2\pi}$ has signified an impressed frequency, it may now signify a frequency of oscillation of the circuit, for a current of frequency $\frac{\omega}{2\pi}$ can exist with zero impressed emf. of that frequency provided the equivalent resistance and reactance for that frequency are both zero. It is then possible to have an im-

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pressed emf. of one frequency $\frac{\omega}{2\pi}$ different from $\frac{\omega_n}{2\pi}$ and yet have the system oscillate in another frequency $\frac{\omega}{2\pi}$ different from the above two frequencies and also different from either frequency $\frac{\omega_1}{2\pi}$ and $\frac{\omega_2}{2\pi}$. It therefore becomes necessary to keep in mind five frequencies.

Referring to equations (34) and (35), oscillation will take place if

(42) $$Z_1 \bar{Z}_2 + M^2 \omega^2 = 0 \quad \textit{General condition of oscillation}$$

where now $\frac{\omega}{2\pi}$ is any frequency instead of the specific frequency of the emf. E_1.

Expanding (42) we get

(43)
$$\left.\begin{aligned} &(R_1 + jX_1)(\bar{R}_2 + j\bar{X}_2) + M^2\omega^2 = 0 \\ \text{or} \quad &(R_1\bar{R}_2 - X_1\bar{X}_2 + M^2\omega^2) + j(X_1\bar{R}_2 + \bar{X}_2 R_1) = 0 \end{aligned}\right\} \textit{General condition of oscillation.}$$

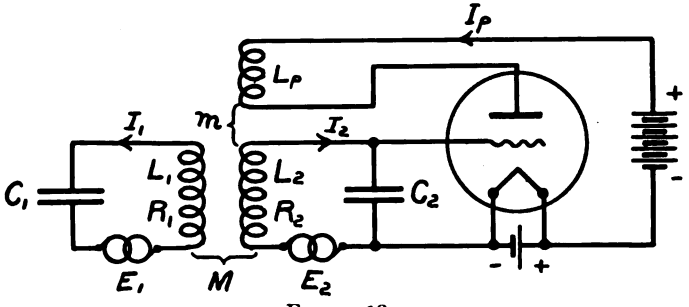

FIGURE 13

The condition of oscillation, therefore, reduces to the double condition that both the real and imaginary parts of (43) must vanish, or

(44) $$\left.\begin{aligned} &R_1\bar{R}_2 - X_1\bar{X}_2 + M^2\omega^2 = 0 \\ &X_1\bar{R}_2 + \bar{X}_2 R_1 = 0 \end{aligned}\right\} \textit{General condition of oscillation.}$$

These equations when solved give expressions previously deduced, that is,

(45-1) $$\bar{R}_2 + \frac{M^2\omega^2 R_1}{R_1^2 + X_1^2} = \bar{R}_{21} = 0 \quad \text{same as (39-3)}$$

(46-1) $$\bar{X}_2 - \frac{M^2\omega^2 X_1}{R_1^2 + X_1^2} = \bar{X}_{21} = 0$$

} *General condition of oscillation.*

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Using the usual notation, equations (45) and (46) may be written as follows:

(45-a) $$\bar{\gamma}_2 = -\frac{\bar{\tau}^2 \gamma_1}{\gamma_1^2 \theta^2 + \left[1 - \frac{\theta^2}{\theta_1^2}\right]^2}$$

(46-a) $$\left[1 - \frac{\theta^2}{\bar{\theta}^2_2}\right] = \frac{\bar{\tau}^2 \left[1 - \frac{\theta^2}{\theta_1^2}\right]}{\gamma_1^2 \theta^2 + \left[1 - \frac{\theta^2}{\theta_1^2}\right]^2}$$

General condition of oscillation.

If θ, the only function containing ω, is eliminated from these two equations, the resulting expression gives the locus of points on the $\theta_1 - \bar{\theta}_2$ plane where equation (42) is true and hence gives the boundary between oscillation and non-oscillation irrespective of the frequency. The elimination yields the expression

(47)

$$[\gamma_1 \gamma_2(\gamma_1 + \gamma_2)\bar{\theta}_2^2 + \bar{\gamma}_2 + \bar{\tau}^2 \gamma_1]\theta_1^4 + [\gamma_2^2(\gamma_1 + \bar{\gamma}_2)\bar{\theta}_2^2 - 2\bar{\gamma}_2(1 - \bar{\tau}^2)]\theta_1^2 \bar{\theta}^2_2 + \left[\frac{\gamma_2}{\gamma_1}(\gamma_1 + \bar{\tau}^2 \gamma_2)\right]\bar{\theta}_2^4 = 0$$

Boundary equation for special case.

This equation is plotted in Figure 14 for $\tau = 0.5$, $\gamma_1 = .4$, and various values of $\bar{\gamma}_2$. Considering any one curve, no frequency $\frac{\omega}{2\pi}$ can make the effective resistance $\bar{R}_{21}$ of the system and the effective reactance $\bar{X}_{21}$ simultaneously vanish for points inside the boundary. The effective resistance of the system viewed from the secondary circuit for points inside the boundary is positive and the system does not oscillate. For each point on the boundary the effective resistance and reactance expressed by equations (45) and (46) vanish for some frequency. For each point outside the boundary, $\bar{R}_{21}$ is negative for some frequencies, one of which makes the reactance $\bar{X}_{21}$ vanish. The system, therefore, oscillates at this frequency, but this condition of negative resistance cannot persist for the current would rise to an infinite value and still have the supply of power greater than the dissipation. The oscillations increase in amplitude until the values of μ and R_p are no longer constant over the cycle and until the average values of these factors have altered to an extent to reduce the effective resistance $\bar{R}_{21}$ for the frequency of oscillation to zero. The stable condition is reached when the altered values of μ and R_p make $\bar{R}_{21}$ and $\bar{X}_{21}$ vanish. The changes in the tube factors do not alter the reactance term nearly as much as they do the resistance term, so that the frequency under these conditions is not

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much different from the frequency which makes the original effective reactance (equation 46) vanish.

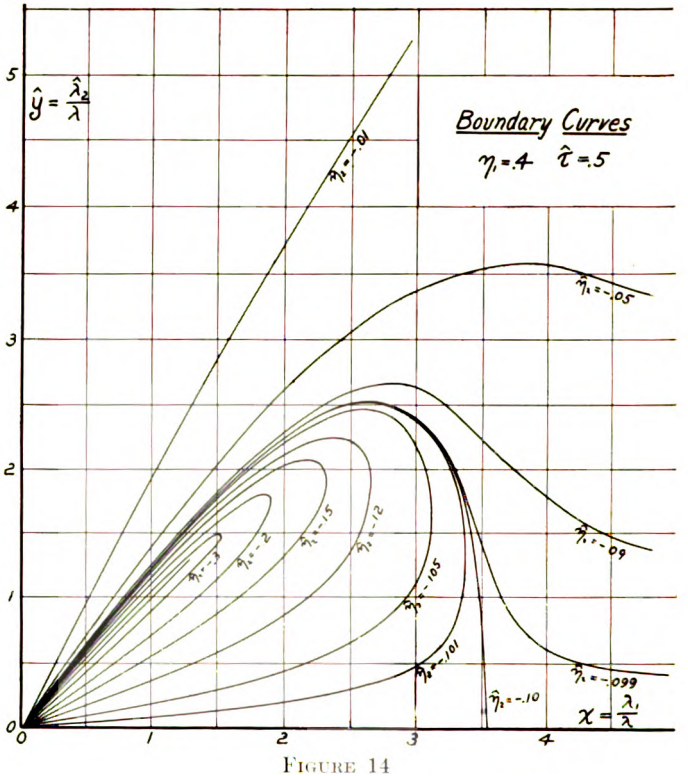

FIGURE 14

Referring to Figure 15, the heavy full line curve marked *boundary* is equation (47) plotted for values of η_1, $\bar{\eta}_2$ and $\bar{\tau}$ given on the figure. This curve represents the boundary between oscillation for points outside and non-oscillation for points within. The coordinates are as indicated, that is, the ratio between the natural wave lengths λ_1 and $\bar{\lambda}_2$ of the respective circuits, to the reference wave length λ_o.

The curves marked Max. 1-2 are equation (46) plotted for various values of θ and mark the loci along which $\overline{X}_{21}$ vanishes for the several values of θ, that is, for different wave lengths. For instance, the Max. 1-2 line for $\theta = 1$ is a line over which $\overline{X}_{21}$ for $\lambda = \bar{\lambda}_o$ is zero, and is the Max. 1-2 line of previous figures over which, when $\bar{R}_2$ was positive, the maximum secondary current was obtained under an impressed emf. of wave length λ_o, and when the order of adjustment of the circuits was first primary

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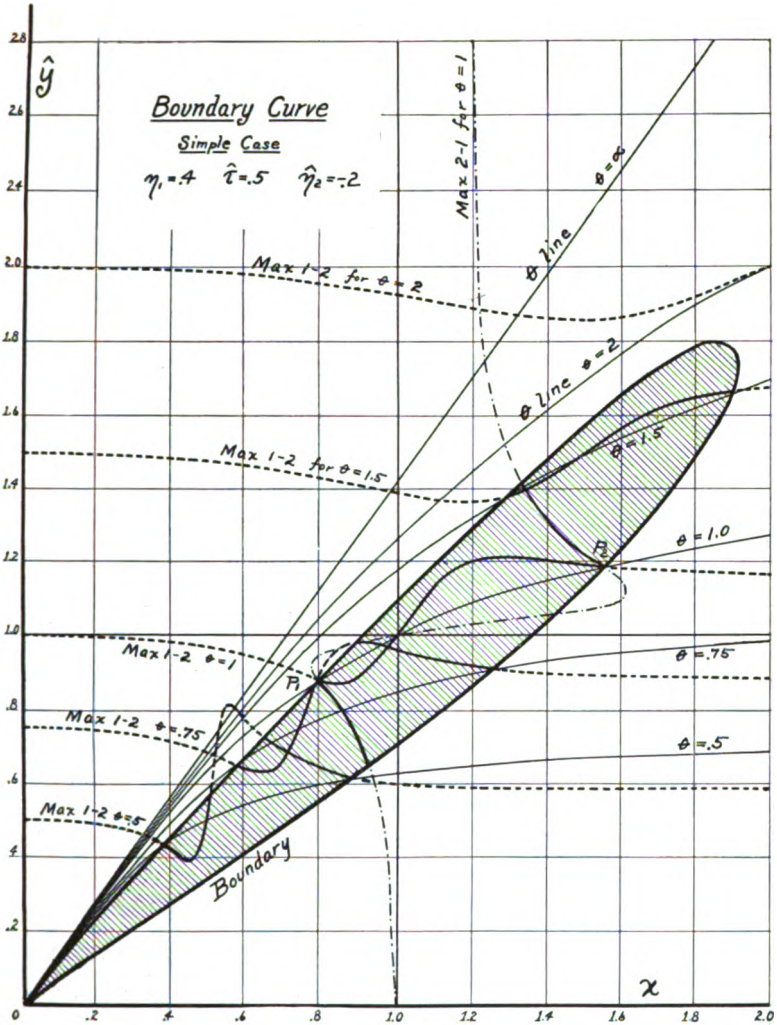

FIGURE 15

and then secondary. Since, in the case represented in Figure 15, $\bar{R}_{21}$ is positive within the boundary, that portion of the Max. 1-2 line for $\theta = 1$ inside this boundary now marks the locus of maximum secondary current when an emf. of wave length λ_o is impressed. The curve outside the boundary approximately gives the line over which the system oscillates with wave length λ_o. The other curves for other values of θ give the analogous curves when $\lambda = \theta \lambda_o$. Inside the boundary the curves give the loci of maximum secondary current under an impressed emf. of wave length

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$\lambda = \theta\lambda_o$, and outside the boundary the curves give the approximate loci of oscillation at wave length $\lambda = \theta\lambda_o$. When a portion of one of the Max. 1-2 lines which is outside the boundary crosses another Max. 1-2 line, there is a condition for which the same adjustments of the circuits satisfy $\bar{X}_{21} = 0$ for the two wave lengths. In this case the oscillation will take place at the longer wave length because then the resistances are less.

If $\bar{\tau}$ be eliminated from equations (45) and (46), then the equation given below results.

(48)
$$\theta^2 = \frac{\bar{r}_1 + \bar{r}_2}{\frac{\bar{r}_1}{\theta_2^2} + \frac{\bar{r}_2}{\theta_1^2}}$$

Locus of max. regeneration points as $\bar{\tau}$ is varied.

If θ equals unity, this equation reduces to equation (40) and gives the locus of the boundary points of oscillation at wave length λ_o as the coupling is varied. For any other value of θ the equation gives the locus of points where oscillation begins for wave length $\theta\lambda_o$ for all values of $\bar{\tau}$.

If the constants of the circuits are as assumed in Figure 15, it is evident that maximum regeneration for wave length λ_o occurs at *two* points where the Max. 1-2 line for $\theta = 1$ cuts the boundary curve.

Figure 16 is similar to Figure 15 except that it is calculated for critical coupling, the numerical value of which for the constants given is $\sqrt{.08}$. In this case the two points of maximum regeneration of Figure 15 have approached and have fused to a single point.

It is now interesting to examine the value of the primary and secondary currents for any impressed frequency $\frac{\omega}{2\pi}$ when the circuits are adjusted to any point of the diagram either inside or outside the boundary curve. The expressions for the primary current taken from equation (34) are

(49-1)
$$I_1 = \frac{E_1}{\bar{Z}_{12}}$$

(49-2)
$$I_1 = \frac{E_1}{\sqrt{\bar{R}_{12}^2 + \bar{X}_{12}^2}} = \frac{\frac{E_1}{L_1\omega_o}}{\sqrt{\left(\frac{\bar{R}_{12}}{L_1\omega_o}\right)^2 + \left(\frac{\bar{X}_{12}}{L_1\omega_o}\right)^2}}$$ or

Numerical value of primary current.

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(49-a)

$$I_1 = \frac{\frac{E_1}{L_1 \omega_0}}{\sqrt{\left[\eta_1 + \frac{\bar{\tau}^2 \bar{\eta}_2}{\bar{\eta}_2^2 \theta^2 + \left[1 - \frac{\theta^2}{\bar{\theta}_2^2}\right]^2}\right]^2 + \frac{1}{\theta}\left[1 - \frac{\theta^2}{\theta_1^2} - \frac{\bar{\tau}^2\left[1 - \frac{\theta^2}{\bar{\theta}_2^2}\right]}{\bar{\eta}_2^2 \theta^2 + \left[1 - \frac{\theta^2}{\bar{\theta}_2^2}\right]^2}\right]^2}}$$

The expression for the secondary current for any frequency $\frac{\omega}{2\pi}$ is from equations (35) and (36)

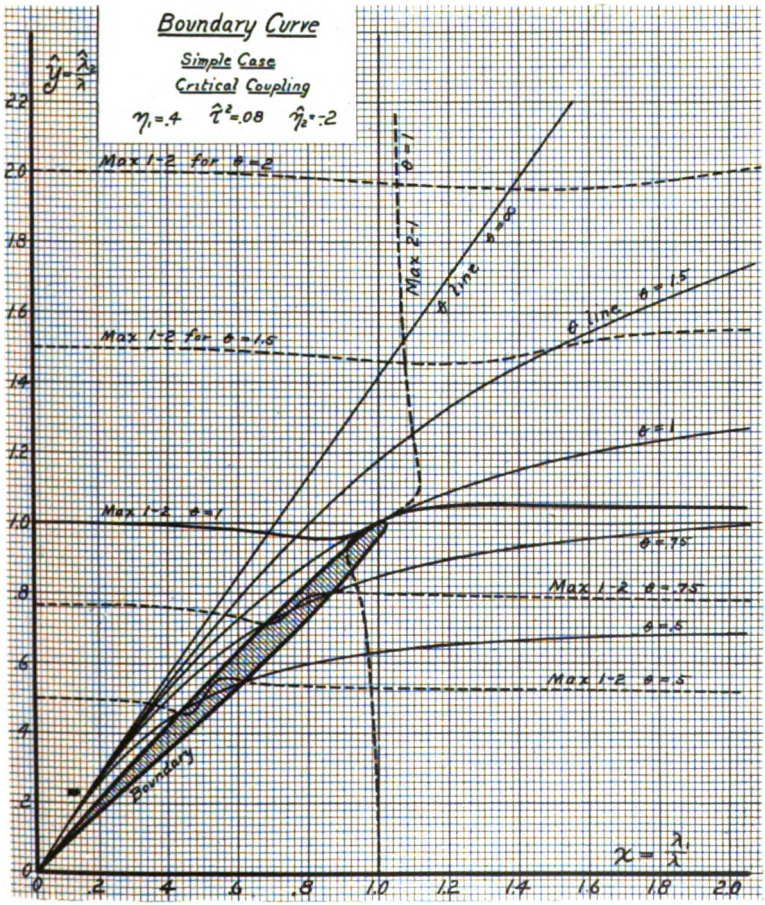
Boundary Curve
Simple Case
Critical Coupling
$\eta_1 = .4$ $\bar{\tau}^2 = .08$ $\bar{\eta}_2 = .2$
$\hat{y} = \frac{\hat{\lambda}_2}{\lambda}$
$x = \frac{\lambda_1}{\lambda}$
Max 1-2 for $\theta = 2$
Max 1-2 for $\theta = 1.5$
Max 1-2 $\theta = 1$
Max 1-2 $\theta = .75$
Max 1-2 $\theta = .5$
θ line $\theta = 2$
θ line $\theta = 1.5$
$\theta = 1$
$\theta = .75$
$\theta = .5$
Max 2-1
$\theta = 1$
Boundary
0 .2 .4 .6 .8 1.0 1.2 1.4 1.6 1.8 2.0
0 .2 .4 .6 .8 1.0 1.2 1.4 1.6 1.8 2.0 2.2

FIGURE 16

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(50-1) $$I_2 = -\frac{jM\omega E_1}{Z_1 \bar{Z}_{21}}$$

(50-2) $$I_2 = \frac{jM\omega E_1}{\sqrt{R_1^2+X_1^2}\ \sqrt{\bar{R}_{21}^2+\bar{X}_{21}^2}} \text{ or}$$

(50-a) $$I_2 = \frac{\frac{\bar{\tau} E}{\omega_o \sqrt{L_1 \bar{L}_2}}}{\sqrt{\left(\theta \frac{R_1}{L_1 \omega_o}\right)^2 + \left(\frac{X_1}{L_1 \omega_o}\right)^2}\ \sqrt{\left(\frac{\bar{R}_{21}}{\bar{L}_2 \omega_o}\right)^2 + \left(\frac{\bar{X}_{21}}{\bar{L}_2 \omega_o}\right)^2}} \text{ or}$$

Numerical value of secondary current

$$I_2 = \frac{\frac{E\bar{\tau}}{\omega_o \sqrt{L_1 \bar{L}_2}}}{\sqrt{\gamma_1^2 \theta^2 + \left[1-\frac{\theta^2}{\theta_1^2}\right]^2}\ \sqrt{\left[\bar{\gamma}_2 + \frac{\bar{\tau}^2 \gamma_1}{\gamma_1^2 \theta^2 + \left[1-\frac{\theta^2}{\theta_1^2}\right]^2}\right]^2 + \frac{1}{\theta^2}\left[1-\frac{\theta^2}{\theta_2^2} - \frac{\bar{\tau}^2\left[1-\frac{\theta^2}{\theta_1^2}\right]}{\gamma_1^2 \theta^2 + \left[1-\frac{\theta^2}{\theta_1^2}\right]^2}\right]^2}}$$

Expressions (49) and (50) are functions of the settings $\frac{\lambda_1}{\lambda_o}$ and $\frac{\lambda_2}{\lambda_o}$ of the two circuits and also functions of $\theta = \frac{\lambda}{\lambda_o}$. The shape of the resonance curve produced by varying λ when the two circuits are set at certain values of $\frac{\lambda_1}{\lambda_o}$ and $\frac{\lambda_2}{\lambda_o}$ can be obtained from equations (49) and (50) by varying θ. This method, of course, gives the exact shape of the resonance curves and is slightly different as already explained from the approximate method described in Part II of taking a cross section thru the space model which has been plotted against x and y. As an example the resonance curve is calculated for the point P_1 of Figure 15. At P_1 as well as at P_2 the system is on the verge of oscillating at wave length λ_o, and hence the system is in a sensitive adjustment for receiving the single wave length λ_o. If the received message is from a radio telephone station the radiated energy consists of a band of wave lengths grouped around the carrier wave λ_o, and it is of interest to obtain the equivalent resonance curve of the system to determine how efficiently the whole radio-telephone spectrum will be received.

Figure 17 is a plot against θ of the components $\frac{\bar{R}_{12}}{L_1 \omega_o}$ and

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$\frac{\bar{X}_{12}}{L_1 \omega_o}$ (see equation (49-2)) of the equivalent impedance of the system of point P_1, Figure 15, as viewed from the primary circuit. For some impressed wave lengths, the system offers a negative resistance to the impressed electromotive force. Under these conditions the wave of wave length λ for which the resistance is negative suffers no absorption in the system; but on the contrary energy is emitted from the system of that wave length, but only when λ is impressed. If the primary circuit is the antenna of a double-circuit receiving system and a wave of length λ is im-

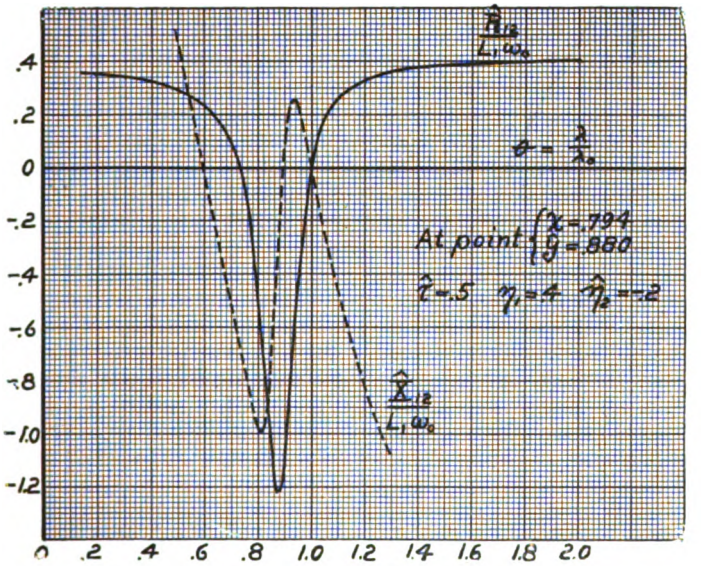

FIGURE 17

pressed for which the resistance is negative, energy will be emitted from the antenna, the phase of the emitted energy being determined by the values of $\bar{X}_{12}$ and $\bar{R}_{12}$. Obviously in some directions around the antenna the direct and emitted wave will reinforce while in other directions the two will tend to cancel. This opens up an interesting field of investigation to determine the effect of one receiving antenna on the distortion of the message received by a neighboring system.

Figure 18 gives plots of $\frac{\bar{R}_{21}}{L_2 \omega_o}$ and $\frac{\bar{X}_{21}}{L_2 \omega_o}$ (see equation (50-a)).

Figure 19 is a plot of the secondary current calculated from equation (50-a).

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The secondary current curve of Figure 19 is, of course, an approximate cross section of the space model along a radial line thru the point P_1. Distances along the radial line are, however, not proportional to θ, but the relation between distance and θ is

$$\text{distance} = \frac{\sqrt{\left(\frac{\lambda_1}{\lambda_0}\right)^2 + \left(\frac{\bar{\lambda}}{\lambda_0}\right)^2}}{\theta} \tag{51}$$

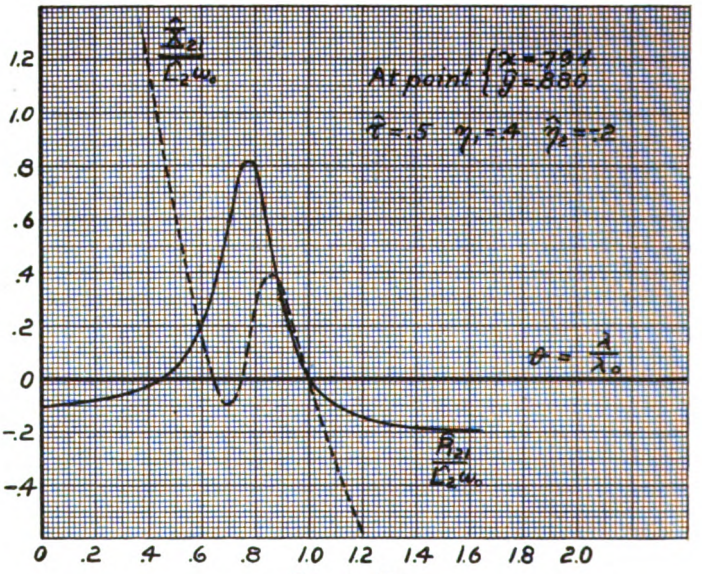

FIGURE 18

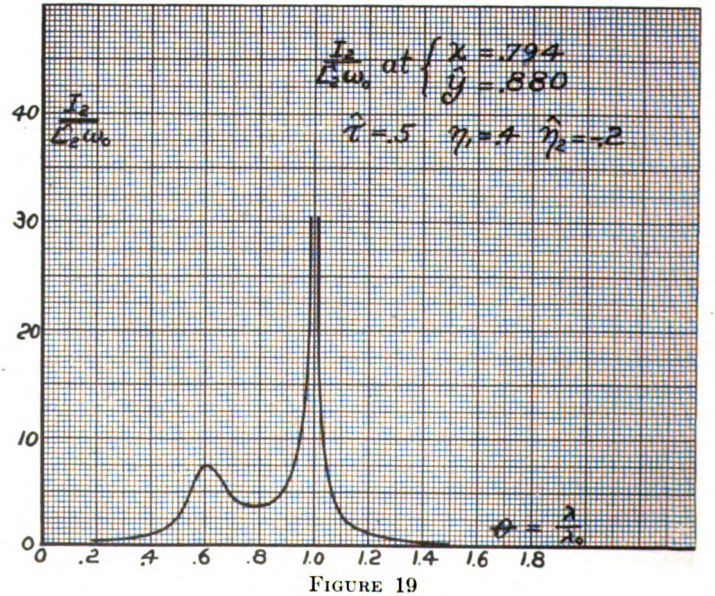

FIGURE 19

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To give a better picture of the cross section of the model Figure 15 is replotted in Figure 20, and along the radial lines the current curves for I_2 are plotted as sections of the space model.

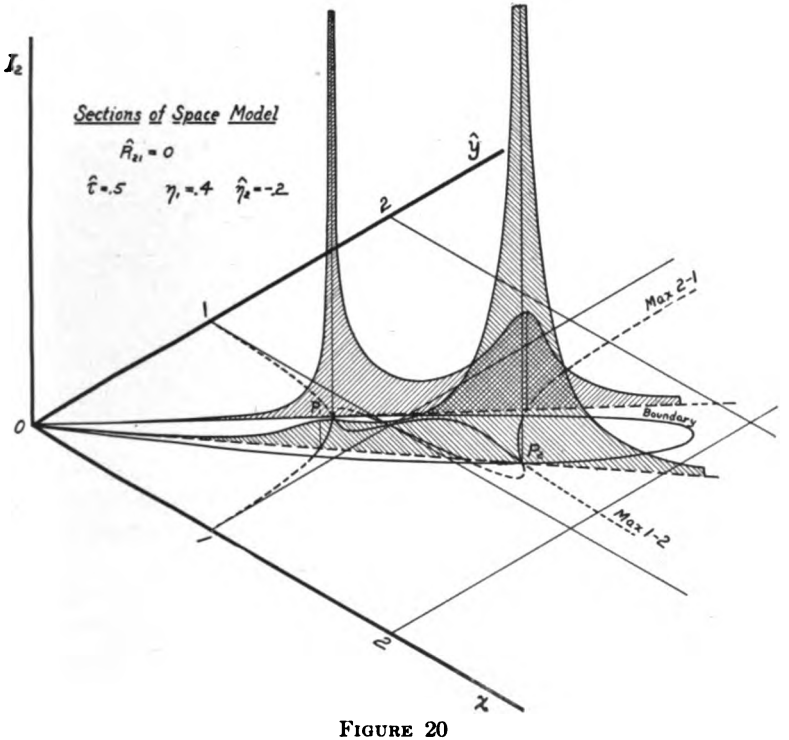

FIGURE 20

Figure 21 shows the values of $\frac{\bar{R}_{12}}{L_1 \omega_o}$ and $\frac{\bar{X}_{12}}{L_1 \omega_o}$ for the point $\frac{\lambda_1}{\lambda_o} = 1$ and $\frac{\bar{\lambda}_2}{\lambda_o} = 1$ on the diagram, Figure 16. This case is for critical coupling and the resistance touches zero in a very small region about the point $\theta = 1$.

CASE 2. COMPLETE CASE OF VARYING REGENERATIVE EFFECT. All of the calculation hitherto considered assumes that the regeneration remains constant as $\frac{\bar{\lambda}_2}{\lambda_o}$ is varied, that is, H defined in equation (25) is assumed constant for any given value of K. This would be true if the tuning of the secondary circuit were accomplished by varying L_2 without changing m. It should also be remembered that for the sake of simplicity all equations and graphs have been expressed in terms of equivalent values of

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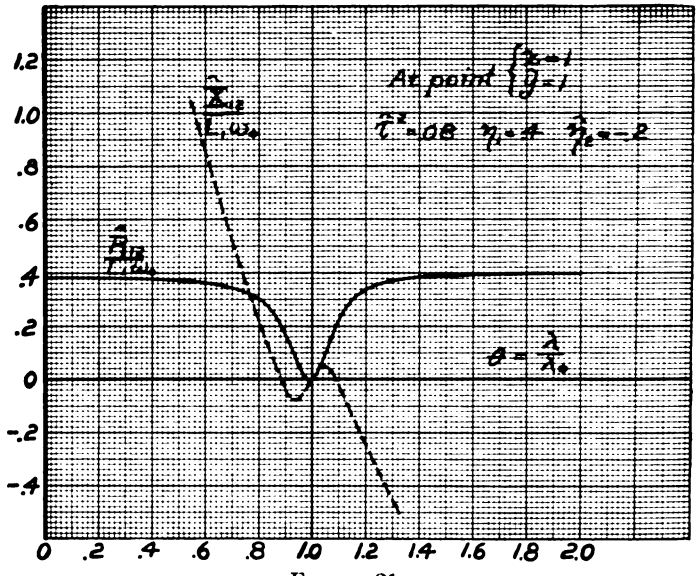

FIGURE 21

$\bar{L}_2$, $\bar{\lambda}_2$, etc., which differ slightly from the actual values of L_2, λ_2, and so on, because of the effect of regeneration on these quantities.

In practice, tuning is usually accomplished by varying C_2. The equations for the ideal case of constant H may be extended to apply to the actual case of practice provided the following assumptions are made which are necessary for simplification.

Assumption 1 $\gamma_1{}^2 \theta^2 \gg 1$

(52) *Assumption* 2 $\mu K \sqrt{\frac{L_2}{L_p}} \cdot \frac{\theta^2}{\theta_2{}^2} \gg K^2$ see equation (25-b)

In actual cases these assumptions cause errors of less than one per cent.

These assumptions lead to the following reduction of equation (25-a).

(53) $$h = \frac{L_2}{L_p} \frac{K \mu \sqrt{\frac{L_2}{L_p}}}{\gamma_1{}^2 \gamma_p{}^2 \theta^2} = \frac{L_2}{L_p} \cdot \frac{s}{\theta_2{}^2}$$ *Approximate value of H*

(54) where $s = \frac{K \mu}{\gamma_1{}^2} \sqrt{\frac{L_2}{L_p}}$

The general conditions of oscillation shown by equations (45-1) and (46-1), now however expressed in terms of actual values instead of equivalent values, are

(45-2) $\bar{R}_{21} = 0$
(46-2) $\bar{X}_{21} = 0$ } *Conditions of oscillation*

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These quantities can now be expressed in terms of S taking account of the fact that C_2 alters the regeneration.

Expanding (45-2) and (46-2) with aid of (53), we have

$$(55\text{-}1)\qquad R_2 - h R_p + \frac{M^2 \omega^2 R_1}{R_1^2 + X_1^2} = \overline{R}_{21} = 0$$

$$(56\text{-}1)\qquad X_2 + h X_p - \frac{M^2 \omega^2 X_1}{R_1^2 + X_1^2} = \overline{X}_{21} = 0$$

Conditions of oscillation, Case 2

These may now be expressed in coefficients as follows:

$$(55\text{-a})\qquad \gamma_2 - \frac{s\,\gamma_p}{\theta_2^2} + \frac{\tau^2 \gamma_1}{\gamma_1^2 \theta^2 + \left[1 - \frac{\theta^2}{\theta_1^2}\right]^2} = 0$$

$$(56\text{-a})\qquad 1 - \frac{\theta^2}{\theta_2^2} + \frac{s}{\theta_2^2} - \frac{\tau^2 \left[1 - \frac{\theta^2}{\theta_1^2}\right]}{\gamma_1^2 \theta^2 + \left[1 - \frac{\theta^2}{\theta_1^2}\right]^2} = 0$$

conditions of oscillations, Case 2

The coupling τ can easily be eliminated from the above equations giving an expression analogous to equation (48). This expression is the locus of points as coupling is varied where oscillation begins for wave length $\theta\,\lambda_o$. The elimination gives

$$(57)\qquad \theta^2 = \theta_1^2 \frac{s(\gamma_p - \gamma_1) - \theta_2^2(\gamma_1 + \gamma_2)}{s\gamma_p - \gamma_1 \theta_1^2 - \gamma_2 \theta_2^2}$$

Locus of max. regeneration points as τ is varied

Substituting equation (57) in (55-a), θ can be eliminated, giving the boundary between oscillation and non-oscillation. This elimination is tedious but yields the following boundary equation.

BOUNDARY EQUATION, COMPLETE SOLUTION

$$(58)\quad \theta_1^4\{\gamma_1\gamma_2(\gamma_1+\gamma_2)\theta_2^4 + [\tau^2\gamma_1 + \gamma_2 + s\gamma_1(\gamma_1\gamma_2 - \gamma_1\gamma_p - 2\gamma_2\gamma_p)]\theta_2^2 + s\gamma_p[s\gamma_1(\gamma_p - \gamma_1) - 1]\} + \theta_1^2\{\gamma_2^2(_1\gamma_1 + \gamma_2)\theta_2^6 + \gamma_2[s(\gamma_1\gamma_2 - 2\gamma_1\gamma_p - 3\gamma_2\gamma_p) - 2 + 2\tau^2]\theta_2^4 + s[s\gamma_p(\gamma_1\gamma_p + 3\gamma_2\gamma_p - 2\gamma_1\gamma_2) + 2(\gamma_p - \gamma_2) - 2\tau^2\gamma_p]\theta_2^2 + s^2\gamma_p[2 - s\gamma_p(\gamma_p - \gamma_1)]\} + \left\{\gamma_2\left[1 + \frac{\tau^2\gamma_2}{\gamma_1}\right]\theta_2^6 + s\left[2\gamma_2 - \gamma_p - \frac{2\tau^2\gamma_2\gamma_p}{\gamma_1}\right]\theta_2^4 + s^2\left[\gamma_2 - 2\gamma_p + \frac{\tau^2\gamma_p^2}{\gamma_1}\right]\theta_2^2 - s^3\gamma_p\right\} = 0$$

Equation (58) is a function of θ_1 and θ_2 and not a function of θ. If, as is done in the figures to be described, equation (58) is plotted to x and y, then, of course, the boundary has a meaning only when the variation of x and y is due to θ_1 and θ_2 and not to θ. So far as the coordinates axes are concerned θ must be assumed equal to unity when considering the boundary equation, but θ will have various values at different points on the boundary,

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the particular value which θ has at any one point determining the frequency at which the circuits oscillate at that point.

Equation (58) is so complex that it gives little idea of the shape of the curve. Three cases are plotted in Figures 22, 23, and 24 for constants given on the plots.

$$I_1 = \frac{\frac{E_1}{L_1\omega_0}}{\sqrt{\left\{\gamma_1 + \left[\frac{\tau^2\left[\gamma_2 - \frac{s\gamma_p}{\theta_2^2}\right]}{2 - \frac{s\gamma_p\theta}{\theta_2^2}}\right]^2 + \left[1 + \frac{s-\theta^2}{\theta_2^2}\right]^2\right\}\left\{\frac{1}{\theta^2}\left[1 - \frac{\theta^2}{H_1^2} - \left[\frac{\tau^2\left[1 + \frac{s-H^2}{H_2^2}\right]}{\gamma_2\theta - \frac{s\gamma_p\theta}{\theta_2^2}}\right]^2 + \left[1 + \frac{s-\theta^2}{\theta_2^2}\right]^2\right]^2\right\}}} \tag{59}$$

$$I_2 = \frac{\frac{E_1}{\omega_0\sqrt{L_1L_2}}\tau}{\sqrt{\left\{\gamma_2 - \frac{s\gamma_p}{H_2^2} + \frac{\tau^2\gamma_1}{\gamma_1^2H^2 + \left(1 - \frac{H^2}{H_1^2}\right)^2}\right\}^2 + \left\{\frac{1}{\theta^2}\left[1 - \frac{\theta^2}{\theta_2^2} + \frac{s}{\theta_2^2} - \frac{\tau^2\left(1 - \frac{H^2}{H_1^2}\right)}{\gamma_1^2\theta^2 + \left(1 - \frac{H^2}{\theta_1^2}\right)^2}\right]^2\right\}}} \tag{60}$$

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Equation (56-1) is the equivalent reactance and corresponds to equation (46-1) of the approximate solution. This equation gives, in the region of non-oscillation, a family of loci of maximum secondary current, each curve being for a constant value of θ. In the region of oscillation each curve gives approximately the locus of oscillation at the particular constant frequency given by the value of θ for that curve. These lines are shown in Figures 22 and 24.

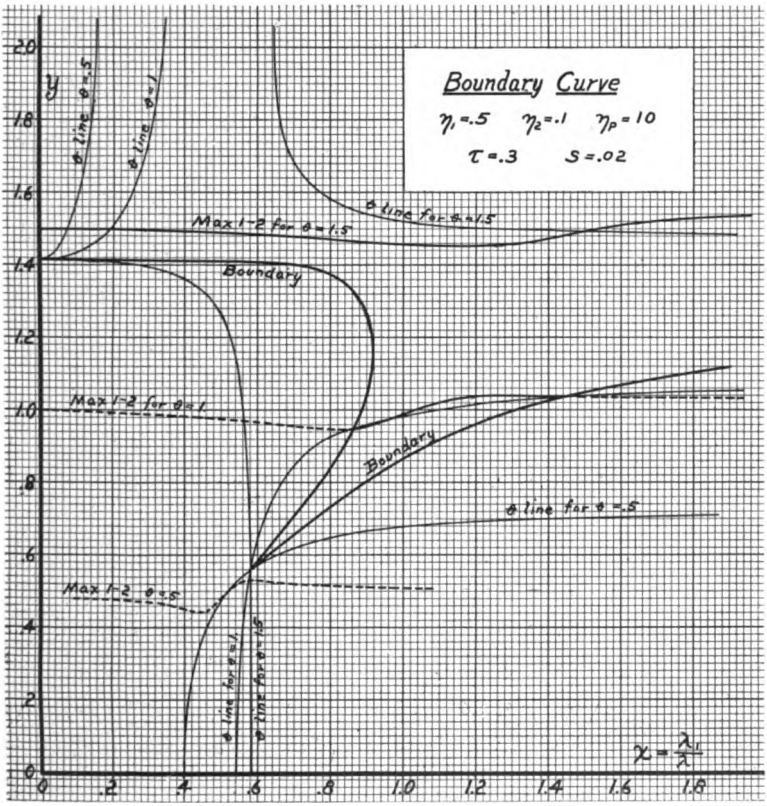

FIGURE 22

Equation (57), which corresponds to equation (48) of the special case solution, is the locus of points of intersection of equation (56-a) and the boundary curve (58) as τ is varied, or in other words, gives the curves over which move the boundary points of oscillation at $\lambda = \theta\lambda_o$ as τ varies. They are indicated in Figures 22 and 24 as θ lines.

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The numerical values of the primary and secondary currents for this complete case are complex. They are given below in terms of co-efficients with the simplification afforded by approximations (52).

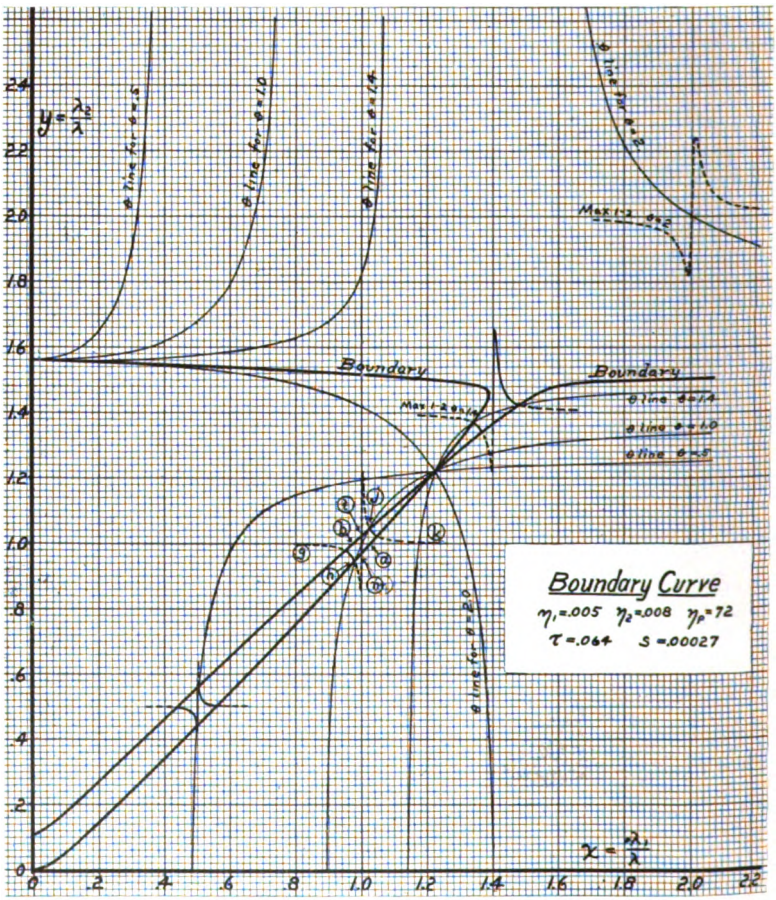

FIGURE 23

Experimental boundary curves are shown in Figures 25, 26, and 27. Figure 25 shows the effect on the boundary curve of varying the coupling between the primary and secondary circuits, other coefficients remaining constant. It is evident that the closer the coupling the broader is the region of non-oscillation. Figure 26 illustrates the effect of changing the amount of regeneration, the arbitrary numbers on the curves being small for the greater regenerative coupling. The boundary curves

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move away from the origin on the 45° line as the regenerative coupling is increased. Figure 27 shows the effect of adding 10 ohms to the primary circuit. A locus of constant oscillation frequency outside the boundary and the maximum-current line inside the boundary are also shown in this figure.

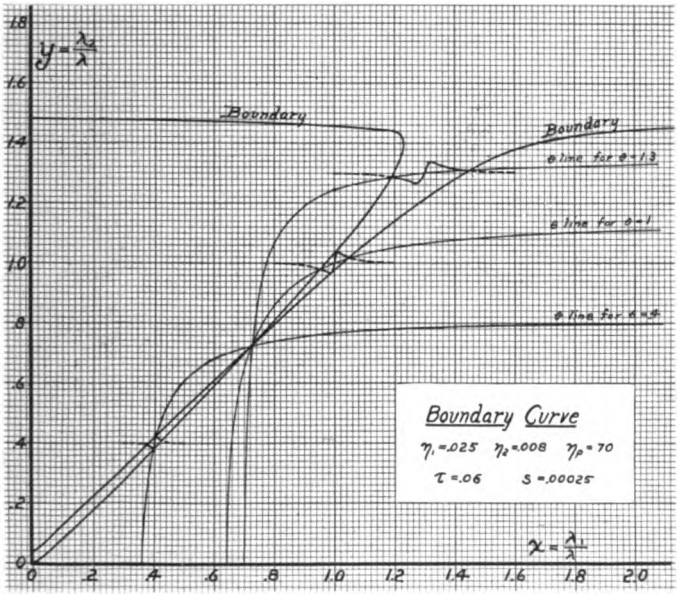

FIGURE 24

The theoretical curves of Figure 22 were calculated for unusually large values of coupling and resistances of the circuits in order to open up the curves to show better their characteristics. The curves of Figure 23 were calculated for the measured constants of the experimental arrangement and should be similar to the curve for no added resistance of Figure 27. The curve of Figure 24 is for a larger value of η_1 and should be similar to the other boundary curve of Figure 27. The values for η_1 and η_2, used in calculating Figure 23, are probably not correct, because the radio frequency resistances of the coils were not measured but estimated from the audio frequency resistances. The agreement of theory and experiment is, however, sufficiently close to confirm the theory.

The curves of Figures 22, 23, and 24, will now be described more in detail. In Figure 22 the boundary curve (equation 58), instead of giving a closed figure extending to the origin as in the

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simple case (Figures 15 and 16) is now an open curve usually coming down to a sharp point at its lowest part. The region above the curve and within the conical shaped part is the region of non-oscillation; the part of the diagram below the boundary curve is the region of oscillation. The intercept on the y axis is the setting for which the secondary circuit as a single circuit starts to oscillate. The boundary curve does not extend to the origin because, as the secondary condenser is decreased, the regenerative tendency rapidly increases, so that for values of the secondary condenser below a certain value, (that is, for the lower extremity or tip of the boundary curve), the system oscillates no matter what the primary circuit setting is.

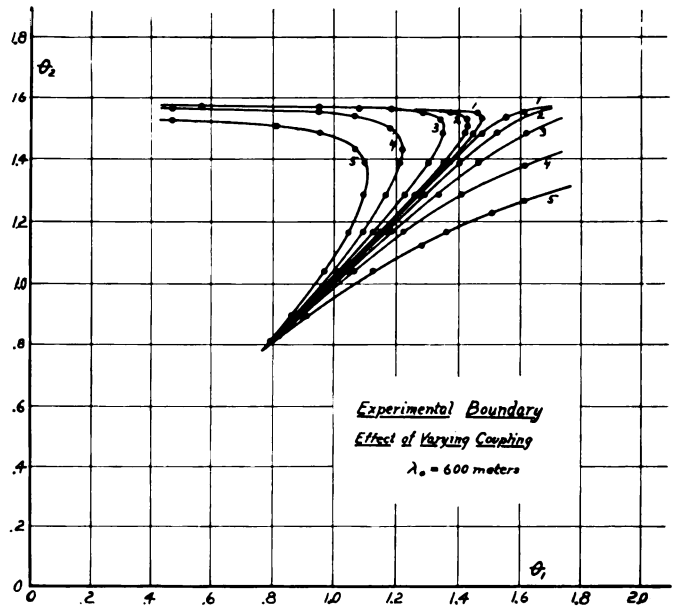

FIGURE 25

The Max. 1-2 lines (equation 56-a) are, as in Case 1, the loci over which X_{21} is zero for various values of θ indicated on the curves. These lines in the region of non-oscillation mark the loci of maximum secondary current when the order of adjustment is first primary and then secondary. They are the max. 1-2 lines of previous figures. In the region of oscillation these curves give approximately the loci of oscillation at wave length $\theta\lambda_o$. The points where these Max. 1-2 lines intersect the boundary curve are points of maximum regeneration at wave length $\theta\lambda_o$ and hence

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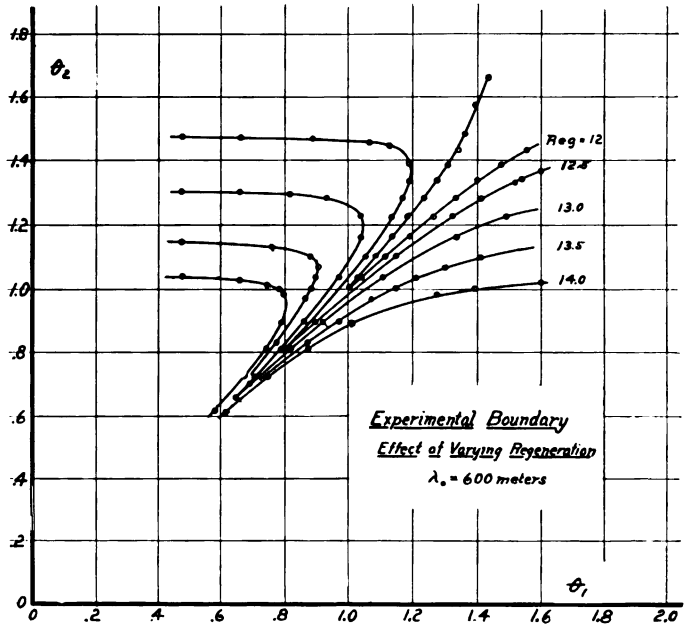

FIGURE 26

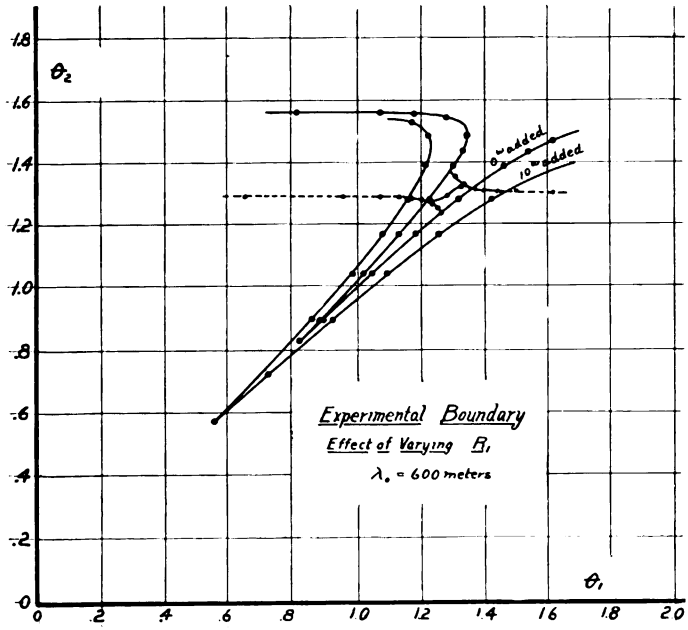

FIGURE 27

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are points of maximum signal strength for that wave length. The θ-line for the same value of θ passes thru these two points of intersection and is the locus over which these two maximum-regeneration points move as the coupling τ is varied. When the Max. 1-2 lines do not intersect the boundary curve there is but one point of intersection of the corresponding θ-line and Max. 1-2 line. The branches of the θ-lines extending upward and above the oscillation region mean nothing and are not included in all of the other figures.

The boundary equation (58) has in some cases as, for example, in the case of Figure 22, imaginary roots for θ_1 for all values of θ_2 below the tip. This is, however, not always so, as shown by the cases of Figures 23 and 24, where values of θ_2 below the tip give real values of θ_1. The boundary equation then encloses a narrow region below the tip which might be supposed to be a region of non-oscillation, but which really is an overlapping of two regions of oscillation, and therefore is a region where the system may oscillate at either one of two frequencies. It is in this region that one gets the familiar click in telephones placed in the plate circuit of a tube generating oscillations in one circuit (the secondary circuit in this case) when another loosely-coupled circuit (the primary circuit) is tuned through the resonant point. For instance, assume that θ_2 is unity and that θ_1 is increased by changing the primary condenser. The coordinate point then moves to the right along the $\theta_2=1$ line of Figure 23. For values of θ_1 between 0 and 1.03 (point *a*), the point is in one region of oscillation, and the frequency of oscillation is given by the value of θ from equation (56-a). This value of θ is given by the upper branch of Figure 28, which is equation (56-a) plotted for $\theta_2=1$. On passing the point *a* ($\theta_1=1.03$) of Figure 23, the representative point passes off one region of oscillation into the second region of oscillation, namely, the right hand region. This change is accompanied by a sudden alteration of frequency of oscillation given by the drop θ from *c* to *d* of Figure 28. As the representative point passes further to the right the value of θ is given by the lower branch of Figure 28 from *d* on toward the right If now θ_1 be decreased, the representative point returns in the second region of oscillation along the $\theta_2=1$ line to point *b* of Figure 23, where an abrupt change of frequency takes place on leaving the second region of oscillation. The variation of θ during this decrease in θ_1 is represented by path *def* of Figure 28.

The enclosed region *cdef* of Figure 28 represents the hysteresis so familiar to experimenters. Sometimes the resonance of two

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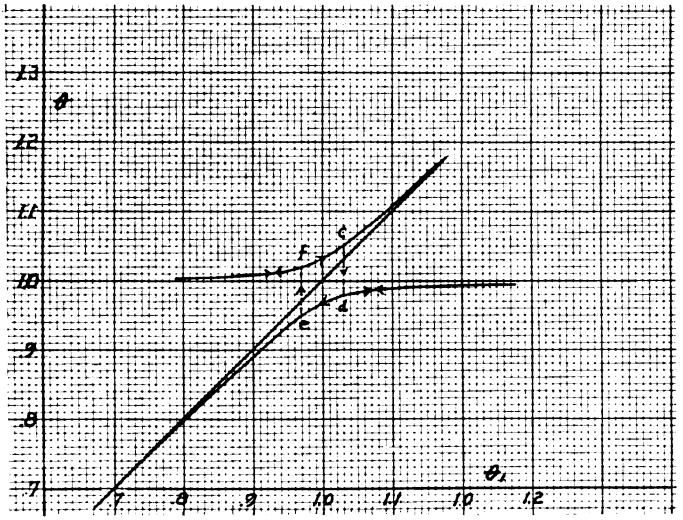

FIGURE 28

circuits is determined by obtaining the average of the values of the condenser settings which give the two clicks. This method is approximate, but gives accurate results if the hysteresis is not great. The hysteresis is reduced by narrowing the region of double oscillation, which in turn can be done by reducing the coupling.

The same sort of hysteresis or abrupt changes in frequency are obtained if the primary condenser is held fixed and the secondary condenser varied. The break points now occur at points l and n of Figure 23. Figure 29, which is equation (56-a) plotted for $\theta_1 = 1$, gives the changes in θ. The path traversed in Figure 29 is *optrsrqpo*.

If, as the primary condenser is increased, the secondary condenser is always adjusted to maintain θ constant at say the value unity, which is to say the system is kept in tune to the incoming wave length $\theta\,\lambda_0$, the representative point moves to the right along curve $g\,h$ of Figure 23. At point h an abrupt change of oscillation frequency takes place. No value of θ_1 and θ_2 between points h and j will make θ equal to unity. For values of θ_1 greater than that for point j the curve $j\,k$ is traversed which lies in the second region of oscillation.

As was pointed out above, there are usually *two* points of intersection of the Max. 1-2 line for $\theta = 1$ with the boundary curve. An experimental determination of the boundary line with the included portions of the Max. 1-2 and Max. 2-1 lines is shown in

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Figure 30. Two cases are shown for two values of coupling between the primary and secondary circuits, the outer boundary being for the greater coupling. The abscissas and ordinates are

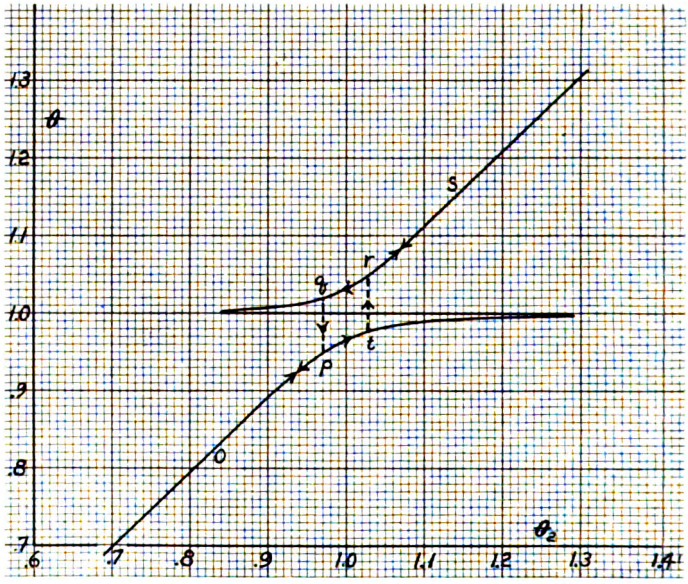

FIGURE 29

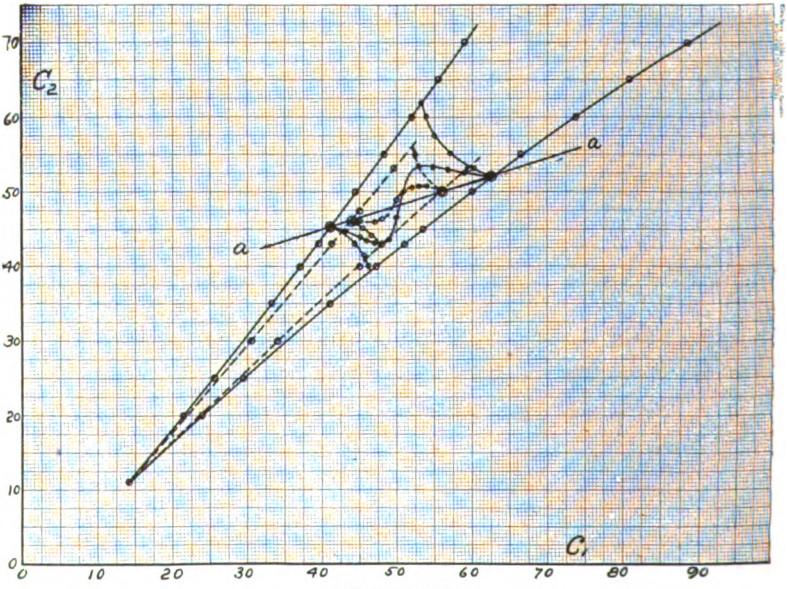

FIGURE 30

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respectively C_1 and C_2. The points of maximum signal strength are shown by large circles.

In tuning a receiving circuit to obtain maximum selectivity there should be but one maximum signal point or one point of intersection of the Max. 1-2 line for $\theta = 1$ with the boundary line. This condition is true when the Max. 1-2 line passes thru the tip of the point and has *no* other intersection with the boundary line.

The tip of the point of the boundary line is usually coincident with the common intersection point of all the θ lines. The coordinates of this point can be obtained from equation (57) and are

$$(61) \qquad \begin{cases} x' = \sqrt{\dfrac{s(\gamma_p + \gamma_2)}{\gamma_1 + \gamma_2}} = \theta_1' \\ y' = \sqrt{\dfrac{s(\gamma_p - \gamma_1)}{\gamma_1 + \gamma_2}} = \theta_2' \end{cases} \qquad \textit{Intersection of } \theta \textit{ line, tip of point.}$$

In dealing with the boundary θ must be taken as unity, so that $x' = \theta_1'$ and $y' = \theta_2'$. The position of the tip of the point is dependent upon s and is *independent* of coupling until the coupling becomes very weak, when the tip suddenly draws away from the point determined by (61), and the boundary line may lose its point and show merely a dip as in Figure 31. An experimentally obtained plot of the position of the tip of the point as indicated by the value of C_2 for various couplings is shown in Figure 32. Plots are given for two values of s or regenerative couplings. The constancy of position of the tip is clearly shown by one curve. For the other curve the regenerative coupling was much reduced, so that the tip of the point is off the plot, but the rising portion touches the axis at C_2 equal to 82, which value is the largest capacity for which the secondary circuit will oscillate alone, and corresponds to the horizontal portion of the boundary line.

Altho the tip of the point may leave the point given by (61), a substitution of (61) in the boundary equation (58) shows that these coordinates satisfy the equation and point (61) must always lie on the boundary curve. In cases similar to that of Figure 36, this point is an isolated portion of the boundary reduced to a single point.

Returning now to Figure 30, it is observed that the max. max. points lie on a line aa, which corresponds to a certain value of $\frac{\gamma_2 \text{ (equivalent)}}{\gamma_1}$. The slope of this line aa shows that $|\bar{\gamma}_2| < \gamma_1$. If the regeneration is increased, $\bar{\gamma}_2$ becomes more negative, and when $-\bar{\gamma}_2 = \gamma_1$, the line aa or its equivalent on the x y plane be-

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comes a 45° line, and the region of no oscillation shrinks practically to a line. Since,

$$\bar{\eta}_2 = \eta_2 - \frac{s\,\eta_p}{\theta_2^2}, \quad \textit{Equivalent } \eta_2 \tag{62}$$

the value of $\bar{\eta}_2$ varies with position on the diagram, so that for only one setting of the secondary circuit with a given value of s does $\bar{\eta}_1 = -\eta$. At only one point, therefore, can the region of no oscillation shrink to zero width, and that is practically at the tip of the point. Above the tip $|\bar{\eta}_2| < \eta_1$. below the tip $|\bar{\eta}_2| < \eta_1$.

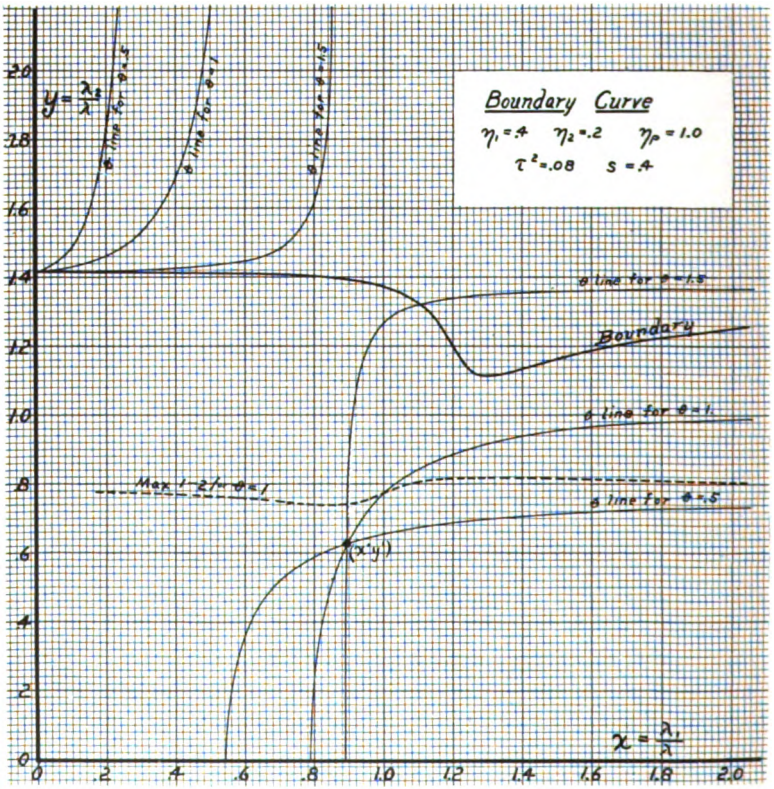

FIGURE 31

These statements are essentially true, but subject to a slight correction as will now be pointed out. The value of θ_2 for the tip of the point is given by equation (61). If this be substituted in (62) the value of $\bar{\eta}_2$ at the tip of the point is

$$\bar{\eta}_2 = -\eta_1 \cdot \frac{\eta_p + \eta_2}{\eta_p - \eta_1}. \quad \textit{Value of } \bar{\eta}_2 \textit{ at tip.} \tag{63}$$

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Since η_2 and η_1 are usually very small compared to η_p, then $\bar{\eta}_2$ is to all intents and purposes equal to $-\eta_1$ at the tip.

The tip of the point can be made to fall practically, but not quite at point $\theta_1=1$, $\theta_2=1$, by making s have the value

$$s=\frac{\eta_1+\eta_2}{\eta_p} \qquad \text{(64)}$$

Approximately correct value of s for tuning.

This is the best value of s for tuning to an incoming wave $\lambda=\lambda_o$, since it makes $-\bar{\eta}_2$ practically equal to η_1. The Max. 1-2 line for $\theta=1$ may, however, have more than one point of intersection with the boundary. Since the width of the boundary curve and the shape of the Max. 1-2 line both depend upon the value of λ, the coupling must now be adjusted so that there is but one point of intersection of the Max. 1-2 line and the boundary, and hence one point of maximum signal strength. This is then the condition of critical coupling.

Analytically the condition of critical coupling can be found by solving for the coordinates of the intersections of the Max. 1-2 line for $\theta=1$, and the boundary line and determining the condition that there be but one point. This can be done by making $\theta=1$ in equations (55-a) and (56-a) and eliminating θ_2. The values of θ_1 for the two points of intersection are given by the following equation

$$1-\frac{1}{\theta_1^2}=\frac{\tau^2 s\eta_p \pm \sqrt{s^2\eta_p^2\tau^4+4\eta_1\tau^2(1-s)[s\eta_p-\eta_2(1-s)]-4\eta_1^2[s\eta_p-(1-s)]^2}}{2[s\eta_p-\eta_2(1-s)]} \qquad \text{(65)}$$

If the radical is zero, there will be but one point of intersection. The condition that the radical is zero and hence the condition of critical coupling is

$$\tau^2=\frac{-2\eta_1}{s^2\eta_p^2}[s\eta_p-\eta_2(1-s)]\left[1-s-\sqrt{s^2\eta_p^2+(1-s)^2}\right] \qquad \text{(66)}$$

Critical coupling

If, as is usually the case, $s^2\eta_p^2$ and s are negligible compared with unity, then critical coupling may be approximately expressed as follows:

$$\tau^2=-\frac{2\eta_1}{s\eta_p^2}[\eta_2-s\eta_p]=-\frac{2\eta_1\bar{\eta}_2}{s\eta_p^2} \qquad \text{(67)}$$

Approximate critical coupling

where $\bar{\eta}_2$ is the equivalent η of the secondary circuit. Equation (67) gives the value that τ must have to give a single point of tuning when $\bar{\eta}_2$ is negative, that is, when $s\eta_p>\eta_2$. If $\eta_2=s\eta_p$ then $\bar{\eta}_2=0$ and $\tau=0$. As $s\eta_p$ is increased, the numerator increases more rapidly than the denominator and τ increases, but if s is made very large, the limit of τ is seen to be $\sqrt{\frac{2\eta_1}{\eta_p}}$, so that for a

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regenerative circuit τ must always lie between zero and $\sqrt{\frac{2\gamma_1}{\gamma_p}}$.

If, now, the value of s from (64) is substituted in equation (67), the value of critical coupling is given as

$$\tau^2 = \frac{2\gamma_1^2}{\gamma_p(\gamma_1+\gamma_2)} \qquad \textit{Critical coupling.} \tag{68}$$

Referring now to Figure 32, the value of critical coupling given

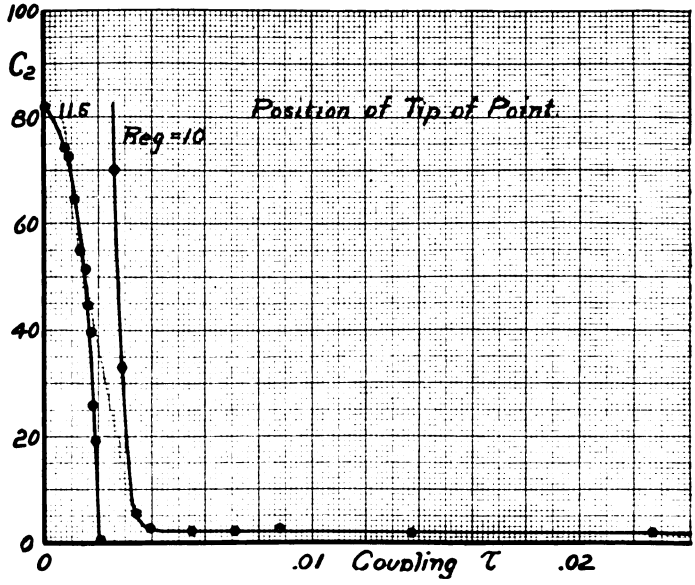

FIGURE 32

by equation (68) is approximately that value when the tip of the point draws away from point (61), or where the curve in Figure 32 starts to rise.

In actual practice it is not always necessary to tune to the very tip of the point, but a slightly closer coupling may be used if the converging portion of the boundary curve is very narrow. Then the two points of maximum regeneration lie so close together that they practically merge into one with slightly broader tuning.

Using the above considerations, a few practical directions for tuning a coupled-circuit receiving system can be given. The value of s, which is determined by the tickler adjustment, must first be set at the correct value given by (64). To find this value approximately, set the coupling at a fairly large value, which is surely above critical coupling, and set the tickler so that the tip

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of the point is near or a little below the wave of the signal being received. The tickler adjustment for this condition represents an upper limit for s. The coupling may now be reduced to the point where a further reduction necessitates a great reduction of the tickler to prevent the system from oscillating. This value of coupling is approximately that given by (68) and represents an upper limit. The values of the tickler and coupling adjustments found above are approximately the theoretically correct values to use, altho practically they represent upper limits and just as good results can be obtained if both are slightly less.

For the next part of the tuning process, increase the coupling a bit and reduce the tickler slightly. Set the primary circuit as near as possible to the incoming wave. Now vary the secondary condenser about the resonant value. The tube will cease to oscillate in a narrow range of secondary capacity somewhere in which the signal should be a maximum. If the range is too wide the coupling should be decreased or the tickler increased. If the position of the maximum signal is not exactly in the center of the non-oscillation range, the primary setting should be slightly altered until, by cut-and-try process, the peak is exactly in the center between the points of oscillation. Now the range may be decreased and the signal increased in strength by decreasing the coupling until the tube nearly oscillates. The final value of the coupling should be about or a little less than the value found above. By following the above procedure the loudest signal with the least distortion and maximum selectivity will be obtained.

A question of considerable interest is the relative selectivity of a coupled-circuit and a single-circuit receiving system. It is well known that the former is much more selective than the latter. An expression can be obtained for the ratio of currents in the tuned circuits connected to the grids of the detecting tubes for the two systems, when the coupled circuit system is set at critical coupling and regenerated to the point of oscillating, and the single circuit is also regenerated to the point of oscillating. Instead of using the complex case for the coupled circuit system, the special solution of Case 1, Part III, which is expressed in terms of equivalent values, is used. This special solution, which is much simpler, is just as applicable as the complete solution, because we are to consider the effects of varying θ, in which case the regeneration remains essentially constant.

The expression for the ratio of currents in the coupled-and single-circuit systems, respectively, when each is set to give the

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strongest signal, can be obtained by dividing equation (50-b) by equation (2-b), remembering, however, that in equation (50-b), θ_1 and $\bar{\theta}_2$ are to be set at unity and $\bar{\tau}$ is to have the critical value given by equation (41-a); and further remembering that in equation (2-b), k is to be adjusted so that the equivalent resistance is zero and that the reactance term stated in equivalent values reduces to $\frac{1-\theta^2}{\theta^2}$. The result of this division is

(69)

$$\frac{I \text{ coupled}}{I \text{ single}} = \sqrt{\frac{\bar{L}_1}{\bar{L}_2}} \cdot \frac{\sqrt{-\eta_1 \bar{\eta}_2 \theta^2}\ \sqrt{\eta_1^2 \theta^2 + (1-\theta^2)^2}}{\sqrt{\bar{\eta}_2^2 \theta^2 (1-\theta^2)^2 + [\eta_1^2 \theta^2 + \eta_1 \bar{\eta}_2 \theta^2 + (1-\theta^2)^2]^2}}$$

It should be remarked in reducing this equation in order to make the comparison of the two systems more just, the primary circuit of the coupled system is assumed to be the antenna circuit of the single-circuit system.

The ratios of voltages impressed on the grids of the detector tubes in the two cases is $\frac{I_2 L_2}{I L}$. This ratio, which is the square root of the ratio of signal strengths in the two systems, is denoted by G, the value of which is given in equation (70).

(70) $$G = \sqrt{\frac{\bar{L}_2}{\bar{L}_1}} \frac{\sqrt{-\eta_1 \bar{\eta}_2 \theta^2}\ \sqrt{\eta_1^2 \theta^2 + (1-\theta^2)^2}}{\sqrt{\bar{\eta}_2^2 \theta^2 (1-\theta^2)^2 + [\eta_1^2 \theta^2 + \eta_1 \bar{\eta}_2 \theta^2 + (1-\theta^2)^2]^2}}$$

(Square root of ratio of signals strengths in coupled and single circuits)

Equation (70) is further simplified by dropping out θ, except in the terms of form $(1-\theta^2)$ on the ground that, for instance, $\eta_1 \theta$ is more nearly constant in practice than η_1, because of the variation of resistance with frequency.

The maximum value of G, called $G_{max.}$ is obtained when θ is equal to unity, and is given below:

(71) $$G_{max} = \frac{\sqrt{-\eta_1 \bar{\eta}_2}}{\eta_1 + \bar{\eta}_2}$$

The ratio of G to G_{max} is probably a better measure of relative selectivity. This ratio is plotted in Figure 33 for $\eta_1 = .10$, and several values of $\bar{\eta}_2$ from 0 to $-.10$, and in Figure 34 for $\eta_2 = .05$ and several values of $\bar{\eta}_2$ from 0 to $-.05$. The values of coupling for the several curves is different in each case, it being determined by equation (41-a). The gain in selectivity through the use of coupled circuits is clearly shown by these figures, in fact, theoretically the selectivity approaches infinity as $-\eta_2$ approaches η_1.

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The values of $G_{max.} \times \sqrt{\frac{L_1}{L_2}}$ for the two values of $\bar{\gamma}_1$ are plotted in Figure 35, which indicates that the coupled system can give a stronger signal than the single circuit when properly tuned.

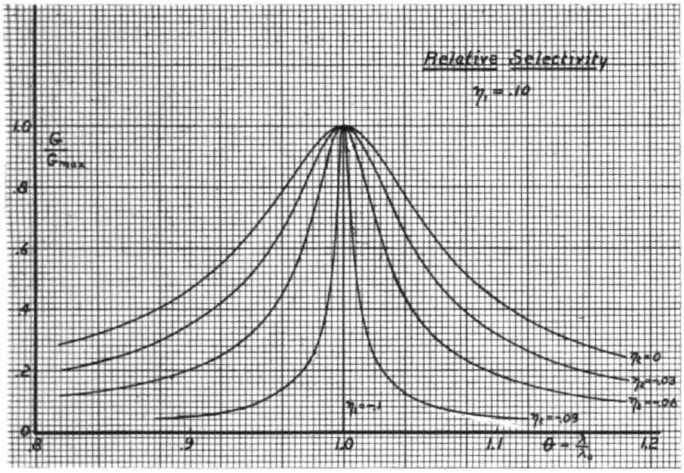

FIGURE 33

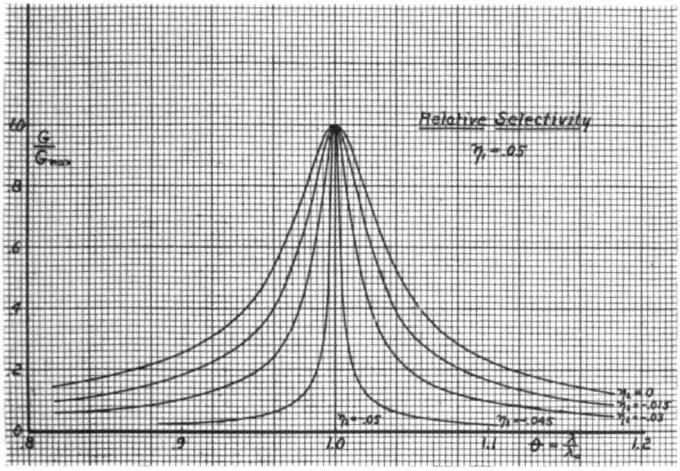

FIGURE 34

Undoubtedly instability reduces somewhat the tremendous theoretical superiority of coupled circuits, both as regards selectivity and signal strength.

Cruft Laboratory, Harvard University,
Cambridge, Mass.

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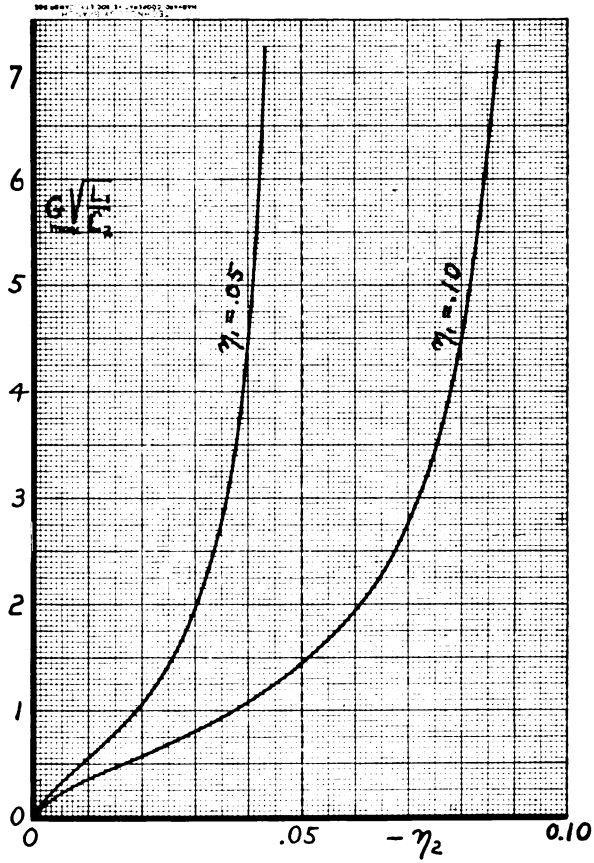

FIGURE 35

SUMMARY: 1—The simple theory of regeneration in a single oscillatory circuit is developed, both when the oscillatory element is in the plate circuit and in the grid circuit of the regenerating tube.

2—The theory of couple circuits having positive resistances is reviewed and illustrated by means of space models for the secondary current. The use of sections of the model for determining graphically and for visualizing the variations of secondary current as the primary or secondary circuits are changed or as the impressed wave length is varied, is given.

The positions of the two peaks of max. max. secondary current, when the coupling is greater than critical, is shown to depend upon the relative values of $\frac{R}{L\omega_0} = \eta$ for the two circuits. When the ratio of $\frac{\eta_1}{\eta_2}$ is unity, the two peaks are symmetrically located on the space model with respect to the primary and secondary axes, otherwise the space model is unsymmetrical.

3—The theory of regeneration in coupled circuits is developed. It is shown that regeneration in the secondary circuit affects only the impedance of that circuit and may make the equivalent resistance of that circuit zero or negative.

The first case taken up assumes the regeneration to be constant, so that the theory for this case is the theory for coupled circuits, one of which has a negative resistance. The boundary equation between oscillation and non-oscillation is derived and plotted.

The second case discussed is the practical case in which the degree of

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regeneration depends upon the size of the secondary condenser. The boundary equation is derived for this case and the curves for several numerical examples plotted. The conditions defining critical coupling and conditions of best tuning of coupled circuits are discussed.

Expressions are derived giving the relative selectivity of coupled circuits and a single circuit receiver, and curves are plotted giving the results in two numerical examples.

NOMENCLATURE

Clarendon Type denotes complex quantities.

Ordinary Type denotes numerical quantities.

$\bar{}$ over a symbol denotes equivalent value.

Equations reduced to forms containing coefficients are indicated by a letter following the equation number, that is, 1-a is the coefficient form for 1-1.

Subscript $_p$ denotes quantities pertaining to the plate circuit of a vacuum tube.

Subscript $_g$ denotes quantities pertaining to the grid circuit of a vacuum tube.

j complex operator numerically equal to $\sqrt{-1}$.

$\mu = -\frac{\partial e_p}{\partial e_g}$ amplification factor of vacuum tube.

$R_p = \frac{\partial e_p}{\partial i_p}$ resistance from plate to filament of a vacuum tube.

R_p may represent the total resistance of the plate circuit, that which is exterior to the tube usually being negligible.

I root-mean-square value of sinusoidal alternating current.

E root-mean-square value of a sinusoidal alternating emf.

R resistance.

L inductance.

C capacity.

Z impedance.

m regenerative mutual inductance.

k regenerative coefficient of coupling.

M mutual inductance between primary and secondary of coupled circuits having inductance L_1 and L_2, respectively.

$\tau = \frac{M}{\sqrt{L_1 L_2}}$ coefficient of coupling between primary and secondary circuits.

ω angular velocity.

$\frac{\omega}{2\pi}$ any frequency, either that of impressed emf. or of oscillation.

λ wave length corresponding to ω.

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$\frac{\omega_o}{2\pi}$ a particular reference frequency.

λ_o wave length corresponding to ω_o.

$\frac{\omega_1}{2\pi} = \frac{1}{2\pi\sqrt{L_1 C_1}}$ the undamped frequency of free oscillation of circuit having inductance L_1 and capacity C_1 (primary circuit).

λ_1 wave length corresponding to ω_1.

$\left.\begin{matrix}\frac{\omega_2}{2\pi} = \frac{1}{2\pi\sqrt{L_2 C_2}} \\ \lambda_2\end{matrix}\right\}$ corresponding quantities for secondary circuit, corresponding to $\frac{\omega_1}{2\pi}$ and λ_1.

$$\theta = \frac{\omega_o}{\omega} = \frac{\lambda}{\lambda_o}$$

$$\theta_1 = \frac{\omega_o}{\omega_1} = \frac{\lambda_1}{\lambda_o}$$

$$\theta_2 = \frac{\omega_o}{\omega_2} = \frac{\lambda_2}{\lambda_o}$$

$$x = \frac{\theta}{\theta_1}$$

$$y = \frac{\theta_2}{\theta}$$

X reactance $= L\omega - \frac{1}{C\omega}$

$$\beta_1 = \left(1 - \frac{\omega_1^2}{\omega_2}\right) = \left(1 - \frac{\lambda^2}{\lambda_2^2}\right) = \left(1 - \frac{\theta^2}{\theta_1^2}\right) = \left(1 - \frac{1}{x^2}\right)$$

$$\beta_2 = \left(1 - \frac{\omega_2^2}{\omega^2}\right) = \left(1 - \frac{\lambda^2}{\lambda_2^2}\right) = \left(1 - \frac{\theta^2}{\theta_2^2}\right) = \left(1 - \frac{1}{y^2}\right)$$

$\gamma = \frac{R}{L\omega_o}$ decrement coefficient at frequency $\frac{\omega_o}{2\pi}$ for circuit denoted by subscripts attached to γ, R, and L.

$\gamma_p = \frac{R_p}{L_p \omega_o}$ decrement coefficient for plate circuit of vacuum tube.

$$s = \frac{k\mu}{\gamma_p^2}\sqrt{\frac{L_2}{L_p}}$$

$$H = \frac{\frac{\mu m}{C_2} - m^2\omega^2}{Z_p^2} = \frac{L_2}{L_p} \cdot \frac{k\left(\frac{\mu}{y^2}\sqrt{\frac{L_2}{L_p}} - k\right)}{\gamma_p^2\theta^2 + 1}$$ coefficient of regeneration.

$$h = \frac{L_2}{L_p} \cdot \frac{k\mu\sqrt{\frac{L_2}{L_p}}}{y^2\gamma_p^2\theta^2} = \frac{L_2}{L_p} \cdot \frac{s}{\theta_2^2}$$ approximate value of H.

$L_2 = L_2 + H L_p$ equivalent inductance of regenerated circuit.

$R_2 = R_2 - H R_p$ equivalent resistance of regenerated circuit.

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$$\bar{\omega}_2 = \frac{1}{\sqrt{\bar{L}_2 C_2}}$$

$$\bar{\lambda}_2 = \frac{2\pi V}{\bar{\omega}_2} \quad V = \text{velocity of light.}$$

$$\bar{\beta}_2 = 1 - \frac{1}{\left(\frac{\bar{\lambda}_2}{\lambda}\right)^2} = 1 - \frac{\theta^2}{\bar{\theta}_2^2}$$

$$\bar{\tau} = \frac{M}{\sqrt{L_1 \bar{L}_2}}$$

$$k = \frac{m}{\sqrt{L_p \bar{L}_2}}$$

$$\eta_2 = \frac{\bar{R}_2}{\bar{L}_2 \omega_2}$$

Z_{12} impedance of coupled circuits looking from primary circuit

Z_{21} impedance of coupled circuits looking from secondary circuit.

G relative selectivity of coupled and single circuit receivers.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

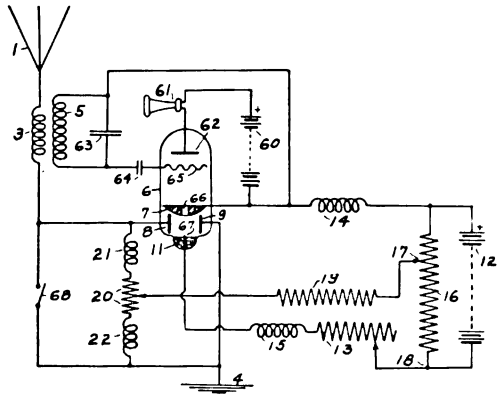
ISSUED MARCH 4, 1924–APRIL 29, 1924

BY

JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, D. C.)

1,485,485—H. G. Cordes, filed September 29, 1919, issued March 4, 1924.



NUMBER 1,485,485—Radio Signaling

RADIO SIGNALING system where a circuit arrangement is provided at the receiving station for reducing the effect of undesirable signals of comparatively short duration and great intensity such as are produced by static disturbances from effecting the receiving circuit. This object is attained by placing a signal sifter in series with the receiving antenna circuit so that signals tending to produce current exceeding a predetermined amplitude cannot pass thru the sifter or are made to pass thru the sifter with increased difficulty as the intensity of the undesirable signal increases. The presence of the sifter in the antenna circuit introduces only a small additional resistance into the circuit for currents of comparatively small amplitude which are produced by the desired signals.

*Received by the Editor, May 15, 1924.

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DIGEST

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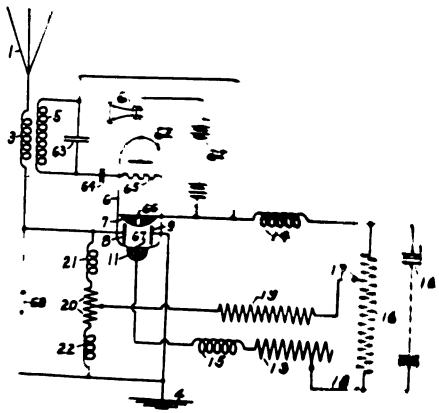
1,455,485—H.

4, 1924.



1,455,485—Radio Signaling

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1,485,524—H. H. Pickron, filed March 9, 1923, issued March 4, 1924. Assigned one-half to W. E. Copp, of Rock Island, Illinois.

CRYSTAL DETECTOR FOR RADIO INSRTUMENTS enclosed in a casing with the mineral supported at one end and a helical telescopic spring located in the other end and adapted for engagement with the mineral.

1,485,773—L. Espenschied, filed September 12, 1921, issued March 4, 1924. Assigned to American Telephone and Telegraph Company.

RADIO CALLING OR SIGNALING from a radio transmitter to remote receiving stations. In brief, the invention consists in using what may be termed the "carrier-off" method, that is, the transmitting station transmits a carrier wave continuously on which a message may from time to time be impressed. When it is desired to call a station, the transmission is suppressed and, as a result, certain operations take place at the receiving station giving some audible or visible signal suitable for calling the attention of the attandant.

1,485,776—J. K. M. Harrison, filed January 3, 1920, issued March 4, 1924.

MARINE SIGNALING APPARATUS for giving a signal as a marking buoy. A galvanic battery is provided with pole elements exposed to the sea water as electrolyte. The energy derived from this battery is utilized for energizing the circuits of a radio transmitter which may be of the buzzer excitation type.

1,486,049—G. B. Spring, filed October 21, 1922, issued March 4, 1924.

RADIO TELEGRAPH AND TELEPHONE INSTRUMENT, consisting of a headset for a radio receiver where the receiver caps are provided with metallic ear pieces contacting with the ears of the operator with connections to the radio receiver whereby the human body may be utilized as the antenna.

1,486,134—H. Gerdien, filed November 30, 1921, issued March 11, 1924. Assigned to Siemens and Halske, Aktiengellschaft.

MEANS FOR TRANSMITTING SIGNALS FOR RADIO TELEGRAPHY, comprising a signal modulation circuit which includes an iron core coil device connected in the antenna circuit. Radio frequency

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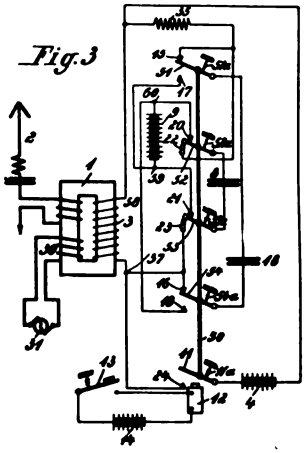
energy is delivered to the antenna thru a coil disposed on the iron core in inductive relation to the antenna coil. A direct current coil is wound on the same core and supplied with low potential direct current. A separate source of direct current is also provided with a circuit arrangement for connecting it to the direct current coil for quickly raising the saturation of the core to the desired point before the low potential direct current supply is connected to the coil whereby the time constant of the system is greatly increased and signals more sharply defined.



NUMBER 1,486,134—Means for Transmitting Signals for Radio Telegraphy

1,486,237—J. A. Fleming, filed November 6, 1919, issued March 11, 1924. Assigned to the Radio Corporation of America, New York.

THERMIONIC DEVICE which consists of vacuous bulb, a filament, a pair of collecting plates and a pair of potential plates, the four plates being arranged substantially symmetrical to and closely surrounding the filament. When the filament is rendered incandescent, negative electricity escapes from the filament and passes across the vacuous space into the plate *C* and then returns by the external circuit passing thru the instrument *F* to the positive terminal of the filament. This current is called the thermionic current. If now the plates *D*, which are called potential plates, are connected to some source of high or low frequency alternating or even direct potential this variation of potential will cause a sudden and marked diminution in the thermionic current, which

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can be observed in the instrument *F* or can be utilized to actuate a relay.

1,486,432—B. Hodgson and S. R. Mullard, filed August 22, 1921, issued March 11, 1924.

SUPPORT FOR FILAMENTS IN THERMIONIC VALVES AND OTHER TUBES, wherein a spring is interposed between the filament and a point on the glass envelope whereby the spring maintains a uniform tension on the filament during abnormal heating of the envelope.

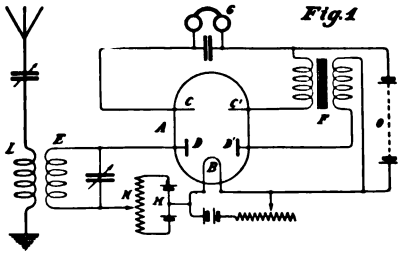
1,486,505—K. W. Wagner, filed June 28, 1923, issued March 11, 1924. Assigned to Radio Corporation of America.

STABLIZING OSCILLATION GENERATORS, wherein a plurality of load circuits are connected with the generator and arranged to maintain the frequency delivered to a work circuit at a constant value.

1,486,506—K. W. Wagner, filed June 28, 1923, issued March 11, 1924. Assigned to Radio Corporation of America.

STABILIZING OSCILLATION GENERATORS, functioning at a particular working frequency. A primary load which is substantially independent of the working frequency is connected to the generator with a filter for excluding secondary or undesired frequencies connected between the generator and the load. A supplemental stabilizing load consisting of an oscillation circuit excluding the working frequency but imposing an increased supplemental load on the generator at nearby frequencies is also connected to the generator whereby the operating condition of the generator tends to remain constant.

1,486,221—E. Berry, filed March 5, 1921, issued March 11, 1924. Assigned to Radio Corporation of America.



NUMBER 1,486,221—Means for Controlling the Flow of Electrons in Electric Discharge Devices

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MEANS FOR CONTROLLING THE FLOW OF ELECTRONS IN ELECTRIC DISCHARGE DEVICES, comprising a tube having a cathode, a pair of anodes, and two plate electrodes located either within or without the vessel and arranged one on each side of the path between the cathode and the anodes so as to apply a potential to the path of the electrons which are thereby caused to fall on the two anodes alternately. The receiving circuit is connected to the electrodes and the detector is connected between the anodes.

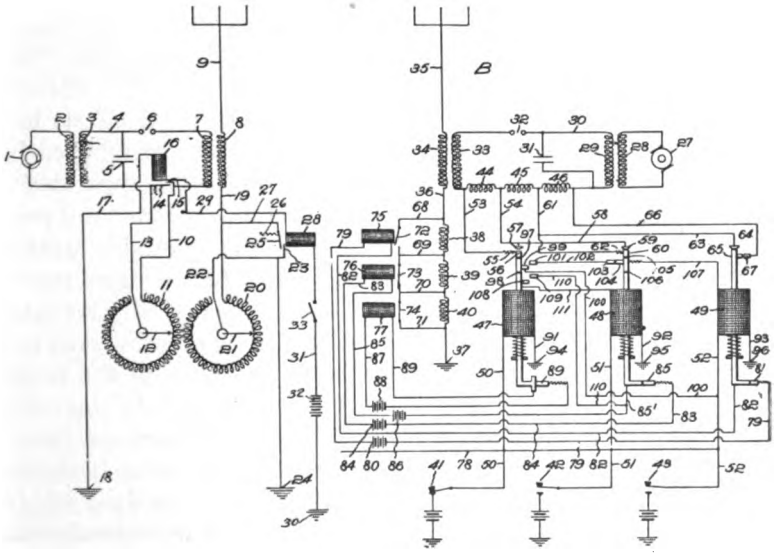
1, 486,885—J. H. Hammond, Jr., filed July 30, 1915, issued March 18, 1924.

RADIODYNAMIC DUPLEX SYSTEM, wherein electromagnetic waves having contrasting characteristics are transmitted for operating a distant control apparatus.

1,486,886—J. H. Hammond, Jr., filed June 3, 1914, issued March 18, 1924.

SYSTEM FOR TELEDYNAMICALLY CONTROLLING MOVING BODIES at the same time that a visible signal carried by such body may also be controlled by radiant energy.

1,486,887—J. H. Hammond, Jr., filed July 13, 1914, issued March 18, 1924.



NUMBER 1,486,887—Electric-Radiant Siren

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ELECTRO-RADIANT SIREN, where signals are transmitted over a varying range of wave lengths. The transmitter contains a pair of variable inductance devices which may be continuously operated to maintain the primary and secondary circuits in resonance over a varying scale of frequencies.

1,487,012—H. Chireix, filed August 29, 1921, issued March 18, 1924.

CALLING ARRANGEMENT FOR SIGNALING, actuating a call indicator in a receiving station in response to a plurality of operations that follow each other in a predetermined cycle. A mechanical system is provided at the receiver including two isochronous pendulums. One of the pendulums is accelerated by the predetermined order of calling signals and then the differential movement of the two pendulums causes an indicating device to be actuated.

1,487,096—L. F. Fuller, filed June 16, 1919, issued March 18, 1924. Assigned to Federal Telegraph Company of San Francisco.

ELECTRICAL CONDENSER of the stack type having a round edge metallic shield at each end thereof. The end plates serve as flux distribution shields increasing the voltage at which corona will form.

1,487,115—J. L. McQuarrie, filed September 15, 1921, issued March 18, 1924. Assigned to Western Electric Company.

INTELLIGENCE SYSTEM for translating electrical effects into mechancial movements, whereby persons who are afflicted by being blind and deaf may read. By the use of the apparatus of this system a person who is both blind and deaf may read print in a convenient and simple manner and with considerable rapidity thru the sense of touch. For accomplishing this, a signal receiving device consisting of a series of electromagnetically actuated levers is arranged so that the free ends of the levers come just beneath the finger tips when the hand is placed on the device thru which the levers pass. The levers are selectively controlled thru the action of selenium cells or other light sensitive means. If a letter of the alphabet, for instance, is interposed between the selenium cell and a source of light, electrical conditions will be set up in circuits containing the selenium cells whereby electromagnets will be operated to actuate its associated lever to con-

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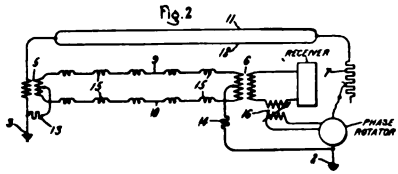
vey to a person who is both blind and deaf, a signal distinctive of a particular character thru the sense of touch.

1,487,298—J. H. Vennes, filed September 5, 1919, issued March 18, 1924. Assigned to Western Electric Company.

METHOD OF AND APPARATUS FOR VIBRATION of electron tubes while connected in a circuit for observing any imperfections in the tube arising out of non-rigid mounting of the electrodes.

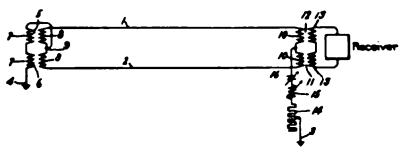
1,487,308—H. H. Beverage, filed January 20, 1921, issued March 18, 1924. Assigned to General Electric Company.



NUMBER 1,487,308—Radio Receiving System

RADIO RECEIVING SYSTEM, where the receiving apparatus may be at a distant point from the receiving antenna. The receiving antenna is a long horizontal uni-directional type with a transmission line for conveying signaling currents from a selected point in the antenna to a distant receiving station. A circuit arrangement is provided for eliminating in the receiving apparatus the effect of currents received upon the transmission line due to the exposure of said line to effect of ether waves.

1,487,339—E. W. Kellogg, filed January 20, 1921, issued March 18, 1924. Assigned to General Electric Company.



NUMBER 1,487,339—Radio Signaling System

RADIO SIGNALING SYSTEM having an antenna for receiving signals from a desired direction to the substantial exclusion of signals coming from other directions. A long horizontal receiving antenna is employed and the received energy conveyed along a transmission line to a distant receiving station located along

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the length of the antenna, from a point in the antenna where the signaling currents received are strongest.

1,487,353—H. J. Nolte, filed September 8, 1921, issued March 18, 1924. Assigned to General Electric Company.

ELECTRON-DISCHARGE APPARATUS, wherein a water cooling system is provided for maintaining the anode at a fairly low temperature. The anode projects into a cooling tank thru which cooling fluid is continuously circulated thru a long spiral heat radiating coil.

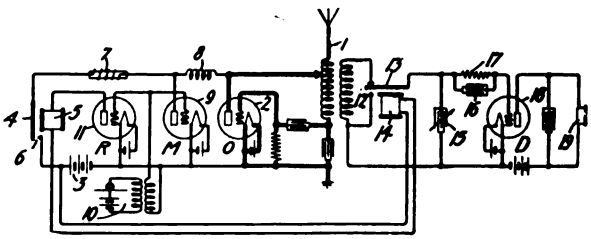
1,487,451—J. F. Farrington, filed October 16, 1922, issued March 18, 1924. Assigned to Western Electric Company, Incorporated.

CIRCUITS FOR ELECTRIC DISCHARGE DEVICES, wherein the cathodes are heated by alternating current. According to the present invention, the disturbances which arise in the output circuit of an electro discharge device in consequence of the alternating current used for heating its cathode may be neutralized by opposing thereto like disturbances produced in the output circuit of a similar discharge device.

1,487,617—D. C. Stoppenbach, filed August 17, 1920, issued March 18, 1924. Assigned to General Electric Company.

ELECTRICAL APPARATUS or condenser built up of alternate sheets of conducting and insulating material. Resilient plates of metal are provided on each end of the stack, which plates are tightly clamped together, uniformly pressing the condenser elements together thruout their area.

1,488,006—R. A. Heising, filed September 29, 1919, issued March 25, 1924. Assigned to Western Electric Company, Incorporated.



NUMBER 1,488,006—Radio Transmission System

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RADIO TRANSMISSION SYSTEM, in which two-way communication may be attained at each station over a single antenna. A circuit is provided for preventing the production of side tone in the receiver at the radio station by energy from the local transmitter. According to this invention, a normally operative receiving set and a normally inoperative transmitting set are connected to the same antenna. Voice or sound operated relays serve, when energized, to render the receiving set inoperative and to render the transmitting set operative. A feature of the invention consists in the provision of a normally oscillating oscillator which serves both for production of oscillations for the transmitter and as a detector. This considerably simplifies the necessary apparatus and makes the control of the system quick acting.

1,488,114—J. H. Hammond, Jr., filed August 24, 1912, issued March 25, 1924.

RADIO TELEGRAPHY AND TELEPHONY, in which the amplitude or frequency or both of the transmitting energy may be periodically varied. A rotary condenser may be interposed in the antenna circuit and driven at a rate where the periodicity of the circuit is varied at a rate above the limits of audibility. The variation is secured in such manner that the time intensity curves shall be peaked and not flattened, whereby a true sinusoidal wave form may be produced at the receiver.

1,488,337—H. Gernsback, filed May 14, 1921, issued March 25, 1924.

ELECTRIC VALVE containing a filament heated to red or white incandescence in contact with the wall of a highly exhausted glass bulb. A remarkably high current is produced between the filament and an outside external electrode. The electronic flow is made to pass between the filament and the outside element, altho these two elements are not in metallic contact. Under the heat of the filament the wall of the glass vessel becomes a conductor which allows the electronic charges to pass. The tube is shown as applied to receiving circuits functioning as a rectifier.

1,488,489—J. C. Gabriel, filed December 23, 1920, issued April 1, 1924. Assigned to Western Electric Company, of New York.

WAVE MODULATING at a radio transmission station. The invention consists in the provision of means for connecting a signaling wave device to a modulating device in two ways, each of which produce modulation and which together cooperate to modulate

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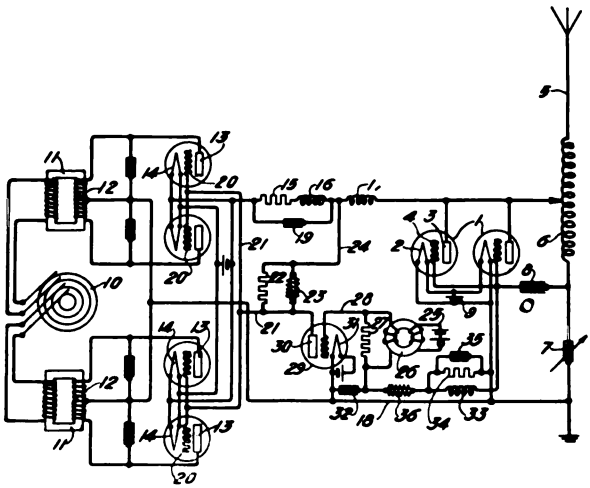
more effectively. The signal-controlling device operates directly thereon thru the agency of an intermediate variable impedance device or system of variable impedance devices such as several discharge tubes arranged to rectify several phases of alternating current and impress the combined rectified current on the modulating device. As a result of the uni-directional conductivity of the variable impedance tubes, the current is variably rectified in accordance with the signals and the resulting waves into which the rectified current is translated are correspondingly modulated in amplitude. Due to the direct action of the signal-controlling device arranged in parallel with respect to the modulator, a further modulation of variation in amplitude of the waves results.



NUMBER 1,488,489—Wave Modulating

1,488,613—G. W. Pickard, filed February 28, 1919, issued April 1, 1924. Assigned to Wireless Specialty Apparatus Company.

VACUOUS ELECTRICAL APPARATUS of the electron tube class wherein desired vacuous condition is preserved without dependance upon a continuously operating vacuum pump. The invention has particular reference to high power transmitting tubes where steel containers are employed in lieu of glass and vacuum maintained despite inherent joints and seals. The metallic container which houses the tube electrodes in a vacuum is itself disposed within an outer vacuous container.

1,488,791—C. Kinsely, filed March 29, 1920, issued April 1, 1924.

SPACE-TELEGRAPH RECEIVING SYSTEM for avoiding the effects

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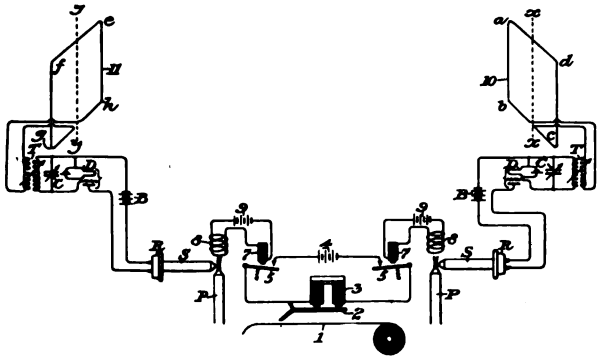
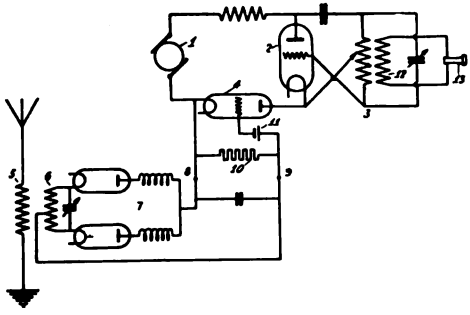
of static disturbances. A pair of loop receivers which may be differently affected by static disturbances are employed connected to Hall relays. The energies thus separately received in a plurality of wave receiving elements are separately utilized to control the operation of a signal indicator circuit so that the indicator in said circuit is actuated only at times when all of said wave-receiving elements are energized. The effects of the strays in several receiving elements are seldom synchronous enough to produce any effect on the indicator.



NUMBER 1,488,791—Space-Telegraph Receiving System

1,489,158—W. Schäffer, filed August 18, 1922, issued April 1, 1924. Assigned to Gesellschaft fur drahtlose Telegraphie, m.b.h.



NUMBER 1,489,158—Arrangement for the Audible Receiving of Undamped Oscillations

ARRANGEMENT FOR THE AUDIBLE RECEIVING OF UNDAMPED OSCILLATIONS, wherein the received energy is used for producing and controlling energy of a different frequency from the received frequency. The frequency thus produced may therefore be kept

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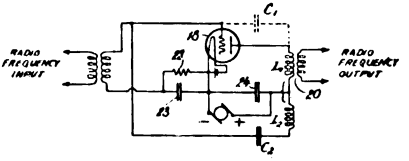
so low that it is audible in a telephone so as to make possible the direct hearing of inaudible incoming radio frequency. An audio frequency generator is provided at the receiving station consisting of an electron tube having a grid for controlling the oscillations in the generator circuit. The receiving circuit causes direct current potential to be impressed on the grid in proportion to the received energy to thereby control the effect of the low frequency generator upon a signal indicator.

1,489,031—J. H. Hammond, Jr., filed March 25, 1914, issued April 1, 1924.

Radiodynmaic System and Method for Avoiding Wave Interference in dynamic control systems. The system contemplates a transmitter of a plurality of differently characterized or contrasting impulses to the first of which only the receiving circuit is responsive and upon receipt of which the electrical constants are automatically changed so that the circuit becomes responsive to the next succeeding transmitted energy upon receipt of which a desired control may be effected. Interference from undesired sources will therefore not affect a control, because the particular combination of frequencies to set a control could probably not be determined by hostile stations.

1,489,228—L. A. Hazeltine, filed December 28, 1920, issued April 1, 1924. Assigned to Hazeltine Corporation of New Jersey.



Number 1,489,228—Method and Means for Neutralizing Capacity Coupling in Audions

Method and Means for Neutralizing Capacity Coupling in Audions, comprising a coil connected between the grid and filament electrode and an auxiliary coil and neutralizing capacity connected in series between the plate electrode and the filament. The auxiliary coil is coupled electromagnetically with the first coil with a coefficient of coupling substantially equal to unity and having a ratio of turns thereto equal to the ratio of the

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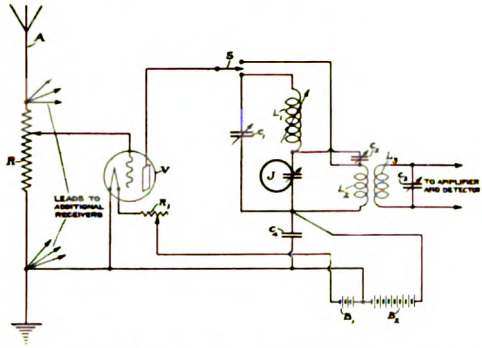
coupling capacity to the neutralizing capacity, said ratio differing from unity.

1,489,287—A. H. Taylor, filed May 29, 1923, issued April 8, 1924.



NUMBER 1,489,287—Receiver of High-Frequency Electrical Signals

RECEIVER OF HIGH-FREQUENCY ELECTRICAL SIGNALS, wherein a multiplicity of separate signal channels may be established on the same antenna. A rejector circuit is interposed between each receiver and the connections with the antenna for eliminating undesired reactions of one receiver upon another.

1,490,165—L. Espenschied, filed September 30, 1919, issued April 15, 1924. Assigned to American Telephone and Telegraph Company.

BALANCED ANTENNA SYSTEM for directive reception so arranged that it will produce a maximum effect upon the receiving apparatus in response to oscillations received in a given direction, but will produce practically no effect upon the receiving apparatus in response to radiations received from the local sending antenna. Another feature in the invention relates to a directive receiving antenna so arranged that its direction of maximum absorbing effect may be rotated at will, without affecting its faculty of producing substantially no effect upon the receiving apparatus in response to radiations transmitted from the local sending antenna.

1,490,198—Q. A. Brackett, filed January 24, 1921, issued April 15, 1924. Assigned to Westinghouse Electric and Manufacturing Company.

SYSTEM OF CONTROL for radio signal transmission where a non-radiating absorbing circuit is provided for the signaling

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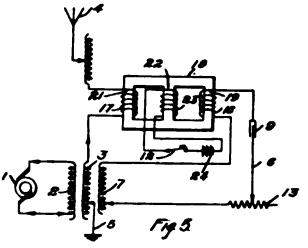
energy between the signal characters. A magnetic core is provided which interlinks inductances connected in the antenna and absorbing circuits. The magnetic characteristics of the core are then changed for producing signals whereby said radiating and absorbing circuits may be selectively rendered effective or non-effective.



NUMBER 1,490,198—System of Control

1,490,958—R. Bown, filed November 23, 1921, issued April 22, 1924. Assigned to American Telephone and Telegraph Company.

FREQUENCY CONTROL SYSTEM, in which a plurality of radio stations are divided into groups of intercommunicating stations, each group having assigned to it a definite frequency range which shall not interfere with the frequency range assigned to any other group. The frequency used in the various groups for signaling is controlled by generating at a master station a fundamental frequency, transmitting said frequency to the stations of the various groups, generating at said station energy to be used for signaling, and controlling by the fundamental frequency of the locally generated energy without controlling the amplitude of the energy.

1,491,288—G. H. Clark, filed May 2, 1919, issued April 22, 1924.

RADIO SIGNALING APPARATUS, including a wave changer system for arc transmitter. A loading coil and a fine adjustment coil are connected in the antenna circuit with switch arms adapted to make contact with said coils. Means are shown for connecting said arms to cause simultaneous movement to provide coarse and fine adjustment.

1,491,341—A. H. Eaves, filed August 14, 1918, issued April 22, 1924. Assigned to Western Electric Company.

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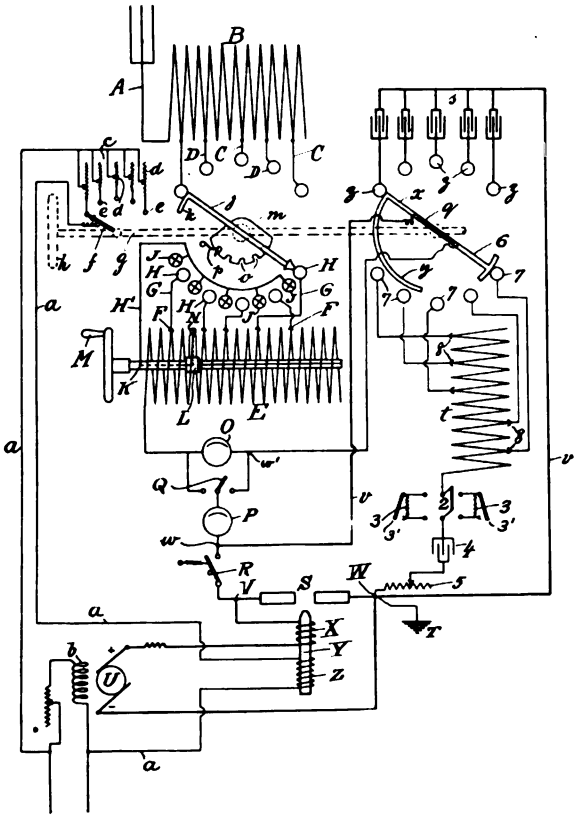
CONDENSER made up of a number of fixed condenser units of various capacities with a switching device whereby the total capacity between a pair of line terminals may be varied in the terms of the smallest capacity.



NUMBER 1,491,288—Radio Signaling Apparatus

1,491,362—H. E. Shreeve, filed February 20, 1918, issued April 22, 1924. Assigned to Western Electric Company.

VACUUM TUBE shell for surrounding the base of a tube. The shell consists of a collar, a plate of insulating material fitted in the outer end of said collar and terminal connections carried by said plate for each of the tube electrodes. The plate is positioned within the collar perpendicular to the axis of the collar.

1,491,372—E. W. F. Alexanderson, filed October 28, 1921,i ssued April 22, 1924. Assigned to General Electric Company.

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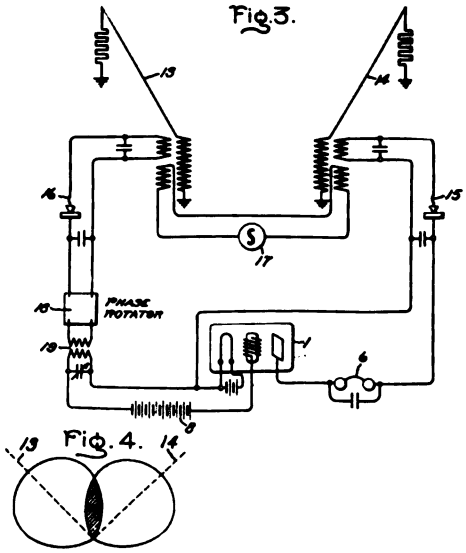

NUMBER 1,491,372—Radio Receiving System

RADIO RECEIVING SYSTEM having a high degree of selectivity. The receiver includes a detector and two circuits associated with said detector upon which received signaling current may be impressed. A circuit is connected to one of these circuits for integrating the phase of the signaling currents flowing therein. An indicator is associated with one of said circuits and separate means associated with the other circuits for causing said indicator to respond only to currents which are in phase with the signaling current. The receiver will then respond only to currents which are in phase with the signaling currents.

1,491,387—P. R. Fortin, filed February 12, 1921, issued April 22, 1924. General Electric Company.

ELECTRON-DISCHARGE APPARATUS, of high power wherein a cooling fluid is circulated around the exterior of the plate electrode of the tube maintaining the electrode at a proper working temperature. The plate electrode is a cylindrical cup-shaped member making a tight joint with the glass envelope of the electron tube.

1,491,405—A. W. Hull, filed March 1, 1921, issued April 22, 1924. Assigned to General Electric Company.

SIGNAL-RECEIVING SYSTEM for the reception of continuous wave signals. A tube circuit is shown in which the potential

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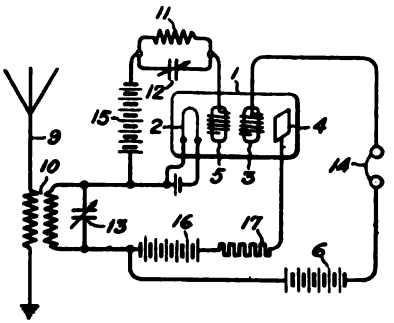
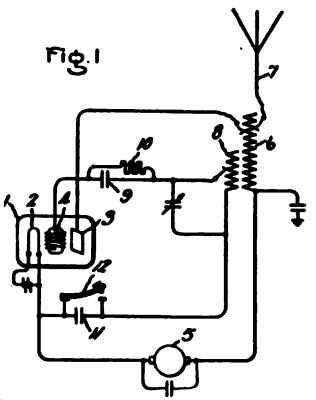
of the grid electrode is varied periodically between suitable positive and negative values at a frequency somewhat different from that of the signaling currents to be detected so that the resistance of the signaling circuit will be varied between maximum and minimum values to vary at an audible frequency. By tuning the circuit betwen cathode and the controlling grid, oscillations will be produced in that circuit and the potential of the grid will automatically be caused to vary periodically at the frequency of the oscillations produced.



NUMBER 1,491,405—Signal-Receiving System

1,491,450—W. C. White, filed February 4, 1920, issued April 22, 1924. Assigned to General Electric Company.



NUMBER 1,491,450—High-Frequency Signaling System

HIGH-FREQUENCY SIGNALING SYSTEM, consisting of an electron tube transmitter wherein signals are produced by a keying circuit connected in the grid of the electron tube. The keying

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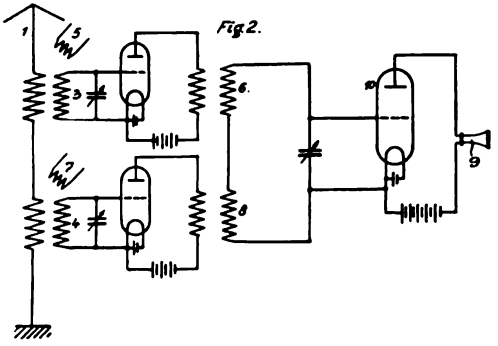
circuit operates to charge the grid to a value sufficiently negative by an accumulation of electrons from the cathode to prevent the production of oscillations so that signaling characters may be produced.

1,491,543—A. Meissner, filed May 3, 1922, issued April 22, 1924.

METHOD OF AND APPARATUS FOR ELIMINATING DISTURBING EFFECTS, by employing two receiving systems each tuned to substantially the same frequency with circuits for modulating the signaling energies of the received effects at different frequencies adapted to produce beats of different frequencies with the signal frequency. A circuit is provided for opposing the differentiated received effects and detecting the resultant of the combined effects.



NUMBER 1,491,543—Method of and Apparatus for Eliminating Disturbing Effects

1,491,772—J. H. Hammond, Jr., filed May 9, 1912, issued April 22, 1924.

SELECTIVE WAVE TRANSMISSION SYSTEM, wherein wave groups are transmitted at a distinctive rate obtained by means of interrupters or variations of wave amplitude. The transmission system includes a circuit inductively connected to the transmission circuit including a source of periodic impulses of a predetermined frequency. These impulses are varied at a frequency different from the first frequency. Another circuit is provided wherein periodic impulses of a still different frequency are produced. At the receiver a plurality of circuits are used tuned to respond to the several frequencies for operating a control by conjoint action of all frequencies.

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1,491,772—J. H. Hammond, Jr., filed April 26, 1912, issued April 22, 1924.

METHOD OF AND SYSTEM FOR SELECTING WAVE TRANSMISSION, for control systems wherein the transmitting station is provided with circuits for emitting a plurality of co-existing series of waves in groups having different group frequencies. The wave groups are emitted in sets having rates different from the wave group frequencies.

1,491,773—J. H. Hammond, Jr., filed May 9, 1912, issued April 22, 1924.

METHOD OF AND SYSTEM FOR SELECTIVE WAVE TRANSMISSION, wherein selectivity in transmission is secured by employing distinctive wave lengths and distinctive wave-group frequencies. A still higher degree of selectivity is secured by transmitting and receiving these waves and wave-groups at a third distinctive rate, which may be obtained by means of interrupters, variations of wave amplitude or in other ways.

1,491,774—J. H. Hammond, Jr., filed May 13, 1912, issued April 22, 1924.

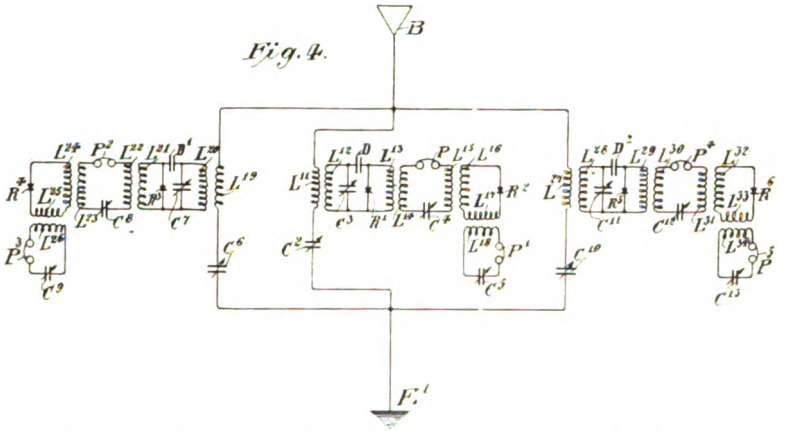

Fig. 4.

NUMBER 1,491,774—Multiplex Radio Telephony and Telegraphy

MULTIPLEX RADIO TELEPHONY AND TELEGRAPHY, where selcetivity is secured by employing distinctive wave lengths and wave-group frequencies or periodic amplitude variations. A still higher degree of selectivity is obtained by using a combination of radio-frequency waves and at least one series of wave groups

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or periodic amplitude variations for the transmission of telephone conversations, and a combination of high-frequency waves and at least two series of wave groups or periodic amplitude variations for the transmission of telegraph signals.

1,491,775—J. H. Hammond, Jr., filed September 29, 1916, issued April 22, 1924.

METHOD OF AND SYSTEM FOR TRANSMITTING AND RECEIVING ELECTRORADIANT ENERGY, for securing secrecy in radio communication. The transmitting circuit shown includes a plurality of rotary spark gaps and a shunted tone circuit whereby a plurality of effective tone signaling frequencies are secured which are selectively received at the receiving circuit to operate any desired control circuit.

1,492,000—H. J. Round, filed August 26, 1920, issued April 29, 1924. Assigned to Radio Corporation of America, New York.

THERMIONIC DEVICE FOR RADIO TELEGRAPHY AND TELEPHONY, in which the anode is situated at a distance from the cathode and is constructed of a pair of independent grid-like devices in close parallel planes. One anode is closer to the cathode than is the other anode. The grid is interposed between the anodes and the positive battery applied to the anode structure.

1,492,321—H. J. J. M. de R. de Bellescize, filed August 29, 1921, issued April 29, 1924.

RADIO SIGNALING SYSTEM, comprising an electron tube apparatus wherein the energy from the detector circuit is delivered to two differential low frequency circuits differently tuned. The two circuits have high time constants with respect to the oscillation frequency and the time constants of the radio frequency receiving circuit.

LIST OF RADIO TRADE MARKS PUBLISHED BY PATENT OFFICE PRIOR TO REGISTRATION

(The numbers given are serial numbers of pending applications)

187,947—"LYTTON RADIO DUPLEX" for radio receiving sets, in ornamental design. Walter Lytton, Incorporated, Chicago, Illinois. Claims use since February 12, 1923. Published March 18, 1924.

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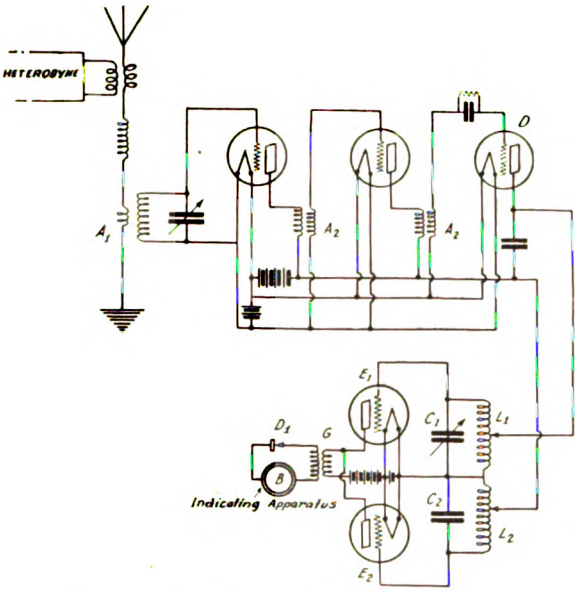

NUMBER 1,492,321—Radio Signaling System

188,673—"SUPERTRAN F M C" for audio frequency transformers. Ford Mica Company, Incorporated, New York. Claims use since October 3, 1923. Published March 18, 1924.

188,897—"RADIO-POWER" for filament heating transformers and parts thereof. The Eagle Carburetor Company, Cleveland, Ohio. Claims use since October 1, 1923. Published March 18, 1924.

188,954—"DUO-REFLEX" for parts for radio receiving sets. Electrical Research Laboratories, Chicago, Illinois. Claims use since April 1, 1923. Published March 18, 1924.

188,955—"ERLADYNE" for parts for radio receiving sets. Electrical Research Laboratories, Chicago, Illinois. Claims use since April 1, 1923. Published March 18, 1924.

188,956—"ERLAFLEX" for parts for radio receiving sets. Electrical Research Laboratories, Chicago, Illinois. Claims use since April 1, 1923. Published March 18, 1924.

188, 957—"TRIPLEX" for parts for radio receiving sets. Electrical Research Laboratories, Chicago, Illinois. Claims use since April 1, 1923. Published March 18, 1924.

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188,970—"PEARLCO" for radio receiving sets. Pearl Radio Corporation, Philadelphia, Pennsylvania. Claims use since March 1, 1922. Published March 18, 1924.

188,972—"RADIONOGRAPH," in ornamental design for radio receiving sets. Radionograph Corporation of America, New York. Claims use since about August 20, 1923. Published March 18, 1924.

188,974—"SEC-TRON" for radio receiving sets. The Sec Tron Company, Cleveland, Ohio. Claims use since May, 1923. Published March 18, 1924.

189,371—Ornamental design of a sound amplifying device for Loud Speakers. Multiple Electric Products Company, New York. Claims use since April 1, 1923. Published March 18, 1924.

163,098—"RADIOPERA" for radio apparatus. Lyon and Healy, Incorporated, Chicago, Illinois. Claims use since about April 1, 1922. Published April 8, 1924.

172,607—"M P M—MILLION POINT MINERAL" for crystal detectors. Million Point Mineral Company, Needles, California. Claims use since October 30, 1920. Published April 8, 1924.

172,769—"WORLD WIDE WIRELESS" in ornamental design for radio apparatus. Radio Corporation of America, New York, New York. Claims use since July, 1921. Published April 8, 1924.

185,289—"DI-LO" in ornamental design for radio receiving apparatus. Fallkill Radio and Electrical Corporation, Poughkeepsie, New York. Claims use since July 1, 1923. Published April 8, 1924.

188,345—"CROSS COUNTRY" for radio receiving sets. United Profit-Sharing Corporation. Claims use since October 1, 1923. Published April 8, 1924.

190,273—"RADIO-MASTER—THE VOICE OF THE WORLD" in ornamental design for radio receiving sets and amplifiers. Henry T. Schiff, Bay City, Michigan. Claims use since October 1, 1923. Published April 8, 1924.

190,274—"VITRODYNE" in ornamental design for radio receiving sets and amplifiers. Henry T. Schiff, Bay City, Michigan. Claims use since October, 1922. Published April 8, 1924.

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190,321—"VITARADIO" for radio receiving sets. Vitanola Talking Machine Company, Chicago, Illinois. Claims use since December, 1922. Published April 8, 1924.

191,082—"MORRADYNE" for radio receiving sets. Morrison Radio-Phone Company, Incorporated. Claims use since January 5, 1921. Published April 8, 1924.

187,654—"THORITE" for loud speakers. Winkler-Reichmann Company, Chicago, Illinois. Claims use since July 29, 1923. Published April 15, 1924.

189,062—"UNI-TUNE" for radio receiving sets. Nassau Radio Company, Incorporated, Brooklyn, New York. Claims use since November 22, 1923. Published April 15, 1924.

189,652—"A-B-ALLEN-BRADLEY" in ornamental design for radio apparatus. Allen Bradley Company, Milwaukee, Wisconsin. Claims use since October 1, 1923. Published April 15, 1924.

190,248—"PYRATEK" for crystal detectors. The Erisman Laboratories, New York, New York. Claims use since December 20, 1923. Published March 18, 1924.

190.297—"PHUSIFORMER" for intervalve coupling transformers. F. R. Hoyt, New York City, assigned to The Wilson Company of New York. Claims use since about November 1, 1923. Published March 18, 1924.

190,686—"HOM-E-OLA" for radio receiving sets. Hoffman-Corr Mfg. Company, New York. Claims use since December 14, 1923. Published March 18, 1924.

190,726—"SUPERTRON" for radio receiving apparatus. Victor Radio Mfg. Company, New York. Claims use since about June 1, 1923. Published March 18, 1924.

191,126—"INDUCTON" for radio receiving sets and parts. Radio Accessories Company, of New York. Claims use since about October 1, 1922. Published March 18, 1924.

191,215—"O. H. VOLTRON" for vacuum tubes. Otto Herschekowitz of Newark, New Jersey. Claims use since November 26, 1923. Published March 18, 1924.

181,886—"HOME-BUILT" for radio receiving coils. G. O. Wilkinson, Philadelphia, Pennsylvania. Claims use since December 20, 1922. Published March 25, 1924. (Not subject to opposition.)

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188,601—Ornamental design for radio apparatus. Mercury Radio Products Company, Little Falls, New Jersey. Claims use since April, 1922. Published March 18, 1924.

191,613—"NEUTROSTAGE" for stage of tuned radio frequency amplification with neutralization of capacity coupling in radio receiving apparatus. Claims use since February 25, 1923. Freed-Eisemann Radio Corporation. Published April 1, 1924.

191,909—"MASTERTONE" for electron tubes. H. G. Porell Company, Boston, Massachusetts. Claims use since December 18, 1923. Published April 1, 1924.

190,380—"K-Q" in ornamental design for radio apparatus. King Sewing Machine Company, Buffalo, New York. Claims use since October 19, 1923. Published April 15, 1924.

190,699—"CATS-EYE" for detectors. Stream-O-Light Lamp Company, Columbus, Ohio. Claims use since March 1, 1920. Published April 15, 1924.

190,709—"RESISTRON" for grid leaks. Temple Instrument Company, Camden, New Jersey. Claims use since September 15, 1923. Published April 15, 1924.

191,362—"ALKADYNE" for radio apparatus, in ornamental design. The Connecticut Telephone and Electric Company, Incorporated, Meriden, Connecticut. Claims use since January 4, 1924. Published April 15, 1924.

191,640—"VOCELESTE" for radio apparatus. The Radio Products Company, Cleveland, Ohio. Claims use since September, 1923. Published April 15, 1924.

192,703—"KANT-BLO" for vacuum tubes. Harry B. Kanter, New York City. Claims use since January 24, 1924. Published April 15, 1924.

182,891—"REDDEN'S" for radio apparatus. (Not subject to opposition.) A. H. Redden, Irvington, New Jersey. Claims use since February 1, 1922. Published April 15, 1924.

192,903—"DURATRAN" for radio transformers. Dubilier Condenser and Radio Corporation, New York. Claims use since about April 28, 1923. Published April 29, 1924.

193,360—"POLYDYNE" for radio apparatus. Service Radio Corporation, New York. Claims use since February 20, 1924. Published April 29, 1924.

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193,361—"POLYMETER" for radio apparatus. Service Radio Corporation, New York. Claims use since February 20, 1924. Published April 29, 1924.

193,362—"POLYDON" for radio apparatus. Service Radio Corporation, New York. Claims use since February 20, 1924. Published April 29, 1924.

193,363—"POLYFORMER" for radio apparatus. Service Radio Corporation, New York. Claims use since February 20, 1924. Published April 29, 1924.

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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 12 AUGUST, 1924 Number 4

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LONG DISTANCE RADIO RECEIVING MEASUREMENTS AT THE BUREAU OF STANDARDS IN 1923*

By
L. W. Austin

(Chief, Radio Physical Laboratory, Bureau of Standards, Washington District of Columbia)

(Communication from the International Union for Scientific Radio Telegraphy)

The results of the signal intensity and atmospheric disturbance measurements for 1923 are shown in the tables of monthly averages. During the past year changes have been made in the wave frequency of Lafayette, and at the same time the antenna at Nauen has been rebuilt so that its antenna current and radiation height have been at times uncertain. These two facts make any comparison with the results of the receiving measurements of the previous year somewhat unsatisfactory. The radiation height of Nauen is now given as 170 meters and the current as 270 amperes, as compared with 150 meters and 380 amperes in 1922. This represents a net decrease of about 25 percent in meter-amperes. Notwithstanding the irregularities in transmitting conditions, the average strength of the Nauen signals for the year is only slightly less than in 1922.

The most remarkable thing to be noticed in connection with the measurements of 1923 is the great increase in signal intensity which has accompanied Lafayette's change in frequency about May 1st, from 12.8 kc. (23,400 m.) to 16.2 kc. (18,500 m.). Thus far no explanation of the increase in the intensity has been offered. Its truth, however, cannot be doubted as the high readings have been noted not only at the Bureau of Standards but also at other American laboratories and at the French army experimental laboratory at Meudon[1] near Paris. The daily records at Lafayette show that there has been practically no change in average antenna current due to the change in frequency, the extremes for 1923 being approximately 400 amperes and 540 amperes with

*Published by permission of the Director of the Bureau of Standards of the U. S. Department of Commerce. Received by the Editor March 13, 1924.

[1] "Onde Electrique," November 25, page 46, 1924

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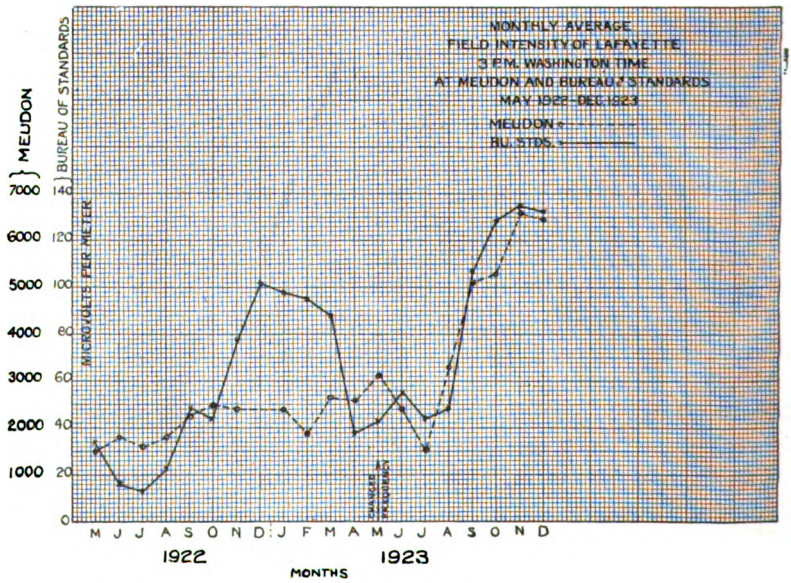
an average of 475 amperes. Such a large increase in strength cannot be due to the suppression of harmonics, nor has there been any change in the antenna system. If the effect were supposed to be due to a change in antenna radiation height, assuming that the height measurement of 170 meters made in 1921 was correct, the present radiation height, as estimated from the increase in intensity observed at the Bureau of Standards and at Meudon, would have to be more than 60 meters above the tops of the towers.

The following figure shows the comparative trend of the monthly averages of the measurements made on Lafayette's URSI signals at 3 P. M., Washington time, at Meudon near Paris, and at the Bureau of Standards. The noticeable thing in the curves aside from the difference in summer and winter values in Washington due to the afternoon fading is the lack of agreement between the Paris and Washington observations before Lafayette's change of wave length and the remarkable agreement shown since that change.



Another fact to be noted in connection with Lafayette's change of frequency is the decrease in its summer afternoon fading at Washington. The unusual weakness of Nauen in the forenoon in December is also worthy of mention. This is probably

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connected with the time of the European sunset and was noticeable, tho in a lesser degree, in other European stations measured. It has been noticed before in the case of Nauen, but in former years the very weak period has more commonly come in November than in December. The European sunset in summer comes in the middle of the afternoon, Washington time, gradually shifting back until during the shortest days of winter the sun sets in Nauen before 10 A. M., Washington time. Observations taken at various times of day thru the different seasons clearly indicate the shift of the weak periods of the European stations, with the changing sunset hour.

In addition to the measurements of Nauen and Lafayette, daily observations were begun in May, 1923, on the Radio Corporation station at Bolinas, California, and on the new French high-power station at Ste. Assise, near Paris. The latter station has a double Marconi type antenna, each half of which is of the same height and length as the antenna at Lafayette, and is so arranged that the two parts can be used simultaneously, one with the call letters UFT employing a frequency of 20.8 kc. ($\lambda = 14{,}400$ m.), and the other UFU with a frequency of 15.0 kc. ($\lambda = 19{,}900$ m.). The observations on Bolinas were discontinued at the end of October on account of extensive repairs and alterations which were being made there, but have now been taken up again.

The frequencies and the approximate antenna currents of the different stations are given in Table III.

As the difficulties of reception in America are almost entirely confined to the summer months, the averages for this period have been collected in Table IV. The degree of fading of the stations, that is, the ratio of the 3 P. M. signal to the 9 A. M. signal is shown in column 5 of the table. Here Lafayette shows the least fading, the ratio of afternoon to morning signal being 0.48. This is an improvement over the value for 1922 in the proportion of 0.48 to 0.35. The Nauen fading is slightly worse than in the year before, being 0.184 against 0.198. It is interesting to note that the fading on the two Ste. Assise frequencies is nearly the same, notwithstanding the considerable difference in frequency, which may indicate that the very marked difference between the fading of Lafayette and Nauen may depend not on the difference of wave length as was surmised last year, but rather on situation or other causes not yet known. The fading on all stations in 1923 continued considerably later than in 1922.

It is worthy of note that Bolinas, tho of nearly the same fre-

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quency as Ste. Assise (UFT) has a much more favorable afternoon fading ratio in Washington, which may be due partly to the difference in conditions along the signal path, and partly to the fact that the sunset effect in California does not coincide in time with the local afternoon absorption around Washington, while the two effects do coincide in the case of Ste. Assise.

The European sunset fading as observed in America is not by any means a daily phenomenon, but occurs irregularly tho often on a number of successive days. It seems to be most noticeable near the time of the longest and the shortest days.

A comparison of the measurements made at the Bureau of Standards with observations made on European stations close to the seacoast indicate that a considerable part of the afternoon fading in Washington is due to absorption taking place within two hundred miles(320 km.).

As has been said before, all conclusions drawn from observations on radio transmission muct be considered strictly tentative until the phenomena have been verified by observations extending over a number of years.

The atmospheric disturbances in the summer of 1923 were less violent than in 1922, but began considerably earlier and continued much later than in that year.

Radio Physical Laboratory,
Bureau of Standards,
March 4, 1924.

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TABLE I

MONTHLY AVERAGES OF SIGNAL INTENSITIES FOR NAUEN AND LAFAYETTE AND ATMOSPHERIC DISTURBANCES IN MICROVOLTS PER METER

1923	Nauen POZ				Lafayette LY			
	A. M.		P. M.		A. M.		P. M.	
	Sig.	Dist.	Sig.	Dist.	Sig.	Dist.	Sig.	Dist.
January	22.3	9	29.6	12	....	19	98.5	25
February	21.2	6	26.3	10	82.9	12	95.8	20
March	34.3	15	30.5	26	81.7	27	89.0	48
April	35.5	63	15.0	152	....	122	38.5	244
May	27.4	97	3.8	306	69.3	129	41.4	370
June	25.3	57	6.3	219	113	77	56.5	383
July	35.2	41	5.5	197	132	56	41.9	232
August	27.8	41	5.0	217	108	49	48.9	255
September	46.2	39	7.9	197	159	46	105	211
October	37.4	23	20.8	78	166	29	129	95
November	24.1	52	20.8	88	146	55	136	95
December	7.7	23	20.2	40	118	29	132	44
Average	28.7	38.8	16.0	129	107	54.2	84.4	168

TABLE II

MONTHLY AVERAGES OF SIGNAL INTENSITIES FOR BOLINAS AND STE. ASSISE IN MICROVOLTS PER METER

1923	Bolinas		Ste. Assise			
	KET		UFT		UFU	
	A. M.	P. M.	A. M.	P. M.	A. M.	P. M.
May	49.8	28.8	43.2	14.2	....	....
June	53.6	21.9	52.3	18.4	101.4	27.0
July	69.4	19.0	56.9	13.4	92.0	24.7
August	55.7	22.0	49.9	10.7	91.4	17.4
September	78.0	32.0	61.9	13.0	112.0	43.4
October	68.8	53.4	52.5	39.0	84.0	66.1
November	....	....	49.4	43.7	109.0	73.2
December	....	....	24.9	33.8	42.0	68.4
Averages	62.5	29.5	48.9	23.2	90.2	45.8

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TABLE III

FREQUENCIES AND APPROXIMATE ANTENNA CURRENTS—1923

	Bolinas	Nauen	Ste. Assise UFT	Ste. Assise UFU	Lafayette
Frequency kc.....	22.9	23.4	20.8	15.0	15.9
Wave length, m....	13,100	12,800	14,400	19,000	18,900
Antenna current, amp.....	420	270	380	475	475

TABLE IV

SIGNAL AVERAGES AND DISTURBANCES DURING AFTERNOON FADING SEASON, JUNE TO SEPTEMBER, 1923

	A. M.		P. M.		Sig. P.M. / Sig. A.M	Sig./Dist.	
	Sig.	Dist.	Sig.	Dist.		A.M.	P.M.
Nauen.....	33.6	44.5	6.2	207	0.18	0.92	0.030
Bolinas.....	64.2	45.0	23.7	208	.38	1.42	.114
Ste:Assise UFT	55.3	44.6	13.8	206	.25	1.24	.067
Ste:Assise UFU	100.5	56.9	28.1	271	.27	1.85	.103
Lafayette.....	129	56.9	62.3	271	.48	2.27	.230

Department of Commerce,
Washington, D. C.

SUMMARY: A summary is given of the signal intensities and strength of atmospheric disturbances at Washington from Nauen, Lafayette, Bolinas, and Ste. Assise stations during 1923.

NOTE: It has been decided to discontinue the bi-monthly publication of the daily observations at the Bureau of Standards on European signals and in atmospheric disturbances, but these will be furnished in mimeographed form to any one making application for them.

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DISTRIBUTION OF RADIO WAVES FROM BROADCASTING STATIONS OVER CITY DISTRICTS*

By

Ralph Bown

and

G. D. Gillett

(Department of Development and Research, American Telephone and Telegraph Company, New York)

With the advent of popular interest in radio broadcasting many of the peculiarities of radio transmission, previously known only to the radio engineer and the radio amateur, have become matters of common experience and broadened interest. The broadcast listener who goes in for receiving distant stations has found that the difference in transmission by day and by night, the enormous fluctuations in night transmission usually termed fading, together with widely varying conditions of natural and man-made interference, lend to his pursuit the sporting aspects of a game of chance. The listener interested in receiving from local stations has learned that two equally powerful stations equally distant from him but in different directions do not necessarily give equal signal strength in his receiver, and again that his friend who lives in a different part of the city gets quite different results. Many of these variations are chargeable to the use of different receiving sets the sensitivity of which varies with frequency in different ways. But after these factors have been cancelled out the fact remains that at short distances from the broadcasting station such that time variations of transmission are inappreciable, or within what might be called the service area of the station, there are space variations or inequalities of distribution which seem to depend on the physical character of the landscape.

There are thus distinguished two kinds of transmission variations; time variations or fading, which are the outstanding characteristics at long distances, and space variations or inequalities of distribution, which are most easily distinguished at short dis-

*Received by the Editor, January 8, 1924. Presented before The Institute of Radio Engineers, New York, January 16, 1924.

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tances where they are not complicated by the presence of time variations. This paper is concerned only with space variations. These are a permanent characteristic of the transmission and lend themselves readily to quantitative studies. To the engineer, a study of irregularities of distribution is valuable not only in disclosing scientific information on radio transmission, but also in providing an empirical basis for the design of radio broascasting distribution systems. We have endeavored to set forth below the results of just such a study of the conditions in the urban and suburban districts of New York City and the City of Washington, District of Columbia.

Theoretical Considerations

Before considering specific cases, it is worth while to discuss very briefly just what the natural obstacles are which lie in the way of theoretically perfect broadcasting distribution.

Looking upon radio broadcasting as a means for distributing intelligence-bearing energy over a given territory, the most desirable result would be to produce the same field strength of radio waves at all points in the territory and to produce little or no disturbance outside that area. This would enable a given receiving set to be used with equal satisfaction anywhere in the territory of the station. Unfortunately radio transmission from a single station is ill adapted to afford such a result. It is by nature a means for throwing out a large amount of energy at one point and allowing it to spread out, diminishing rapidly at first and then more gradually, until finally at considerable distance it becomes unstable or too attenuated to be discernible.

Not only is the field strength reduced with increasing distance on account of the spreading of the energy over a larger and larger circle, but it is also reduced in addition by the fact that some of the radiated energy is transformed into heat by electrical losses in the transmission media and therefore lost to radio uses.

If the surface of the earth were flat and of uniform electrical characteristics, the falling off in field strength of the waves due to spreading of the energy and losses in the earth might be expressed as a law in mathematical form. Several such expressions have been developed from experimental data and are useful in many cases. In the usual case, however, the earth's surface is irregular in form or material and corresponding transmission irregularities are introduced. The characteristics of the earth's

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surface which affect radio transmission may be roughly classified under three headings as follows:

1. Areas of different electrical constants—fresh water, salt water, dry land, wet land, rock, snow, and so on.
2. Differences of elevation—hills, valleys, mountains, and the like.
3. Absorbing structures—man-made buildings, towers, or other structures many of which have resonance characteristics producing selective absorption.

It is almost hopeless to expect that we shall ever be able properly to take account of such varied factors as these in any transmission formula. We may, however, thru the collection of experimental data hope to obtain a fairly clear picture of how they affect transmission and be able to make allowances for them in engineering problems and calculations.

Experimental Methods

Fortunately the means for taking experimental data of this kind are already at hand in the form of radio field-strength meas uring sets of the kind described in a recent paper before The Institute of Radio Engineers on "Radio Transmission Measurements," by Bown, Englund, and Friis.[1] One of the small portable type of short wave measuring sets described in that paper was used in obtaining the data herewith presented. In order to be able to cover ground rapidly with this outfit, it was transported and used in an automobile as shown in Figure 1.

In our measurements at New York and Washington, schedules were arranged with certain stations so that a carrier frequency of constant amplitude would be sent out while observations were being taken. During some of the periods this carrier was modulated or, more strictly speaking, two intelligence-bearing side bands were transmitted in addition to the constant carrier. This was immaterial from the measuring standpoint, because the measuring set measures substantially only the carrier.

It is obvious that taking any considerable number of points to the square mile would be very laborious. The data contained in this paper represent the results of about two months of field work during which the automobile carrying the measuring set traveled 3,000 miles (4,800 kilometers). An endeavor was made, therefore, to reduce the amount of measuring work by taking the observations at points selected in advance as most likely to show

[1] "Proceedings of The Institute of Radio Engineers," volume 11, 1923, number 2, page 115.

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up the effects being investigated. In many places the data are very sketchy, but they serve as a guide in picturing phenomena which are more accurately delineated in other sections where the data are more complete.

The observations in each city group were made during daylight hours, during the same season of the year, and in most instances under uniform weather conditions. Relatively, the measured values of field strength are believed to be correct within a few percent. The absolute accuracy of the values does not enter into this discussion, so the possible 10 or 15 percent constant error in the calibration of the measuring set used is of no direct

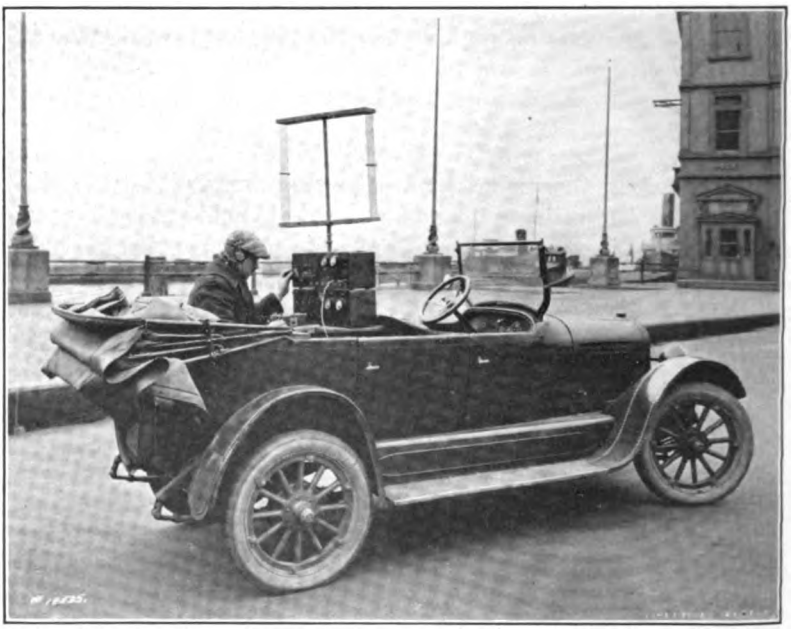

FIGURE 1

importance. Where observations were made in settled areas, as open locations as possible were chosen, for instance, in parks, or squares and in the centers of wide streets. Such points tend to represent general conditions as distinct from the very local conditions which may obtain on the roofs of or inside of particular buildings. No doubt if the study were carried to sufficient refinement, each building or group of buildings would be found to have its own little distribution pattern.

In order to facilitate an orderly consideration of the data,

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we have presented them below in such sequence as to bring out, first, the separate effects of the three classes of transmission factors previously mentioned, and, second, the complications into which combinations of the factors lead.

Open Country Transmission

The first measurements to be considered were made around Washington, District of Columbia, in the summer of 1923. The broadcasting station WCAP of the Chesapeake and Potomac Telephone Company was used as a transmitting station for the tests. The transmitting frequency was 640 kilocycles (wave length 469 meters). As shown on the map, Figure 2, observations were taken on two lines leading out of the city, one to the southeast and the other to the northwest. The points at which

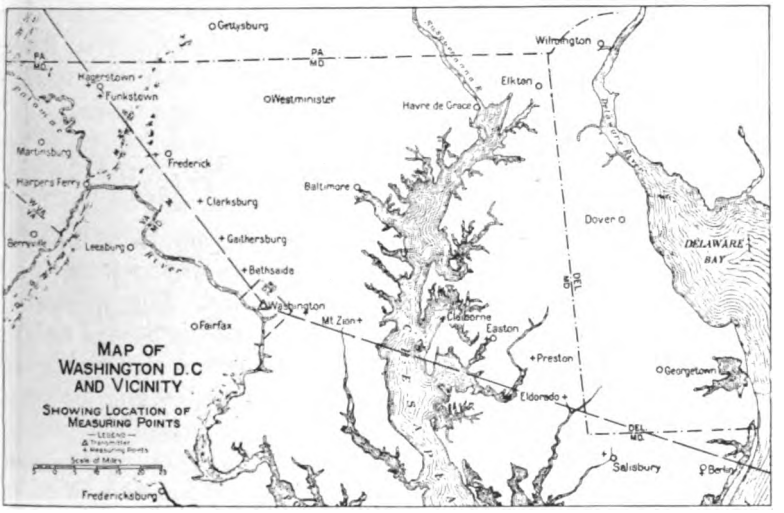

FIGURE 2—Map of Washington, District of Columbia, and Vicinity showing Location of Measuring Points

measurements were made are marked. In Figure 3 are given curves showing plots of field strength against the distance from the transmitting station. The smooth upper curve *A* shows the falling off which would be obtained under the ideal condition of simple spreading, the field strength being inversely proportional to the distance. The calculation of the ideal curve requires an assumption or measurement of the effective height of the transmitting antenna. In the present case, the effective height was calculated from field strength measurements made at a dis-

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tance of about eight wave lengths from the transmitting station.[2]

Curve *B* is for the southeast line. A profile of the land and water corresponding to this line is given just below at *B'*. It starts out over rolling country and the field drops rapidly. About 25 miles (40 km.) out it strikes Chesapeake Bay, and the next 30 miles (48 km.) over water is characterized by a flattening of the field strength curve, due to lowered attenuation. Then, going over land again, the curve drops more rapidly. The alternating steep and flat portions bring out clearly the effect of the different electrical characteristics of the land and brackish water over which the radio waves pass.

Curve *C*, with its corresponding profile of the landscape *C'*, is for the northwest line, which is entirely over land. The land increases in elevation and irregularity as the distance from the city increases. The field strength falls off at first much the same as in Curve *B*. Farther out, as the foothills of the Blue Ridge Mountains are reached, the rate of falling off increases until a low point is registered in the shadow of a mountain range.

A simple theory suffices to explain this shadow effect. If the ground is considered to be a fairly good conductor, the penetration of the waves will not be great, and there will be but little transmission directly thru the mountain. As the waves pass the crest and the ground falls away, the bottom of the wave front is stretched and the unit field strength reduced. Simultaneously the wave front is given a forward tilt. Since the flow of energy is perpendicular to the wave front, this means that energy is being fed down from above. The effect of the feeding down of energy is gradually to erase the shadow, as is brought out by the rising of Curve *C* after the mountain has been well passed. What occurs is probably very similar to the diffraction of light, altho it should be noted that the dimensions of the obstructions are small relative to the wave length. It is probable that if a hundred or more points had been taken, closely spaced along this northwest line, it would be continually rising up and down in minor undulations reflecting the shadows of the smaller hills and mountains. The seven points shown serve to give a general indication of the average rate of attenuation and to make more prominent the one large shadow caused by a natural land elevation.

[2] For a brief discussion of this method of determining the effective height, see paper on "Radio Transmission Measurements," by Bown, Englund. and Friis, "Proceedings of The Institute of Radio Engineers," volume 11, number 2, page 117

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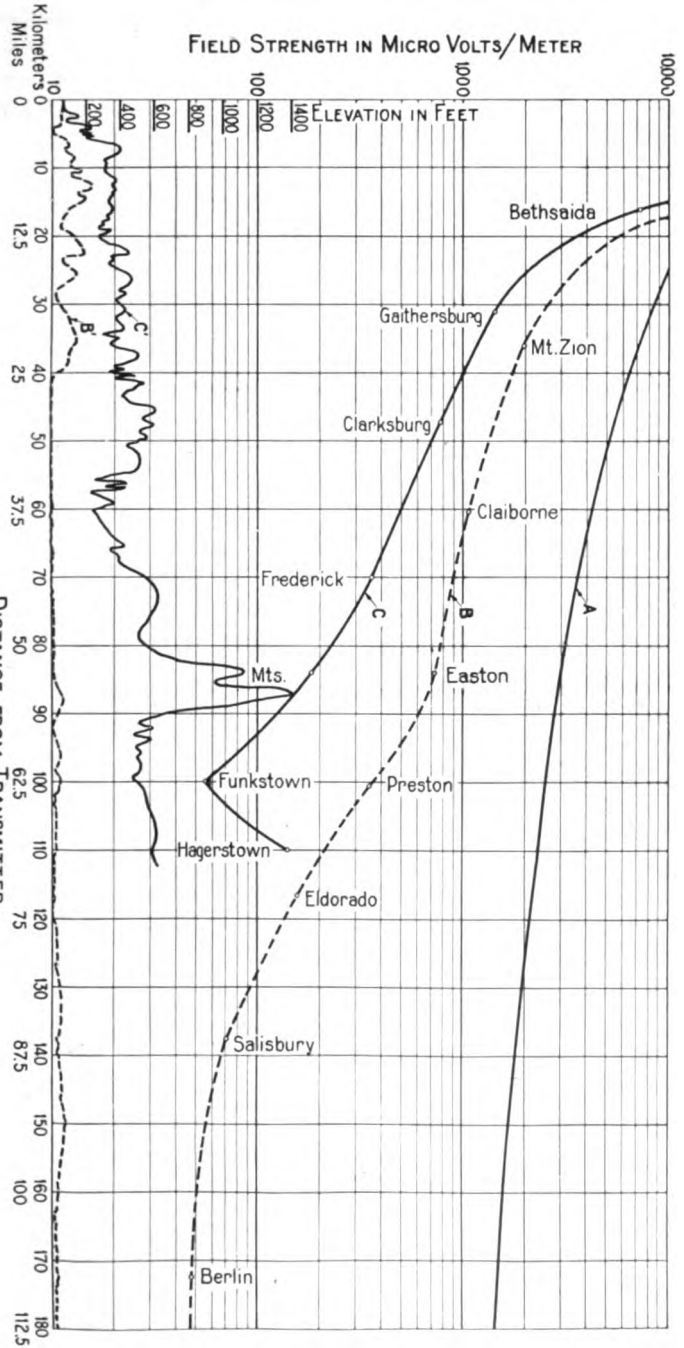

FIGURE 3—Plot showing Effect of Topography on Radio Transmission

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An effort was made to classify the data taken in the country districts outside the city of Washington along the two radial lines just discussed and along other radial lines in other directions, with the object of obtaining an attenuation factor or formula by means of which the transmission over the soil of Maryland and Virginia could be calculated. The wide scattering of the values caused this to be abandoned as impractical. During the analysis there was calculated for each individual point the attenuation factor α which, inserted in the familiar transmission formula, caused it to give the correct field strength for that particular point. The formula is

$$\text{Field strength} = 377 \frac{h_s I_s}{\lambda d} \varepsilon^{-\frac{\alpha d}{\sqrt{\lambda}}} \text{ micro-volts per meter,}$$

where $h_s I_s$ is the antenna radiation constant in meter-amperes, and λ and d are the wave length and distance, respectively, both in kilometers. The values of α obtained by this process ran all the way from 0.012 to 0.035, with a rough average at about 0.022.

One method of analysis applied to the curves of Figure 3 consisted in fitting attenuation factors to sections of the curves which are uniform and correspond to transmission over uniform territory. For instance, the attenuation factor for the section of Curve *B* between Preston and Salisbury is 0.028. Thoe lan is low but dry and sandy. The rolling upland of more mist soild from Gaithersburg out to Frederick (Curve *C*) has an attenuation factor of 0.009. The over-water transmission of Curve *B* between Mt. Zion and Easton shows a factor of 0.0025. This is relatively quite close to the 0.0015 factor of the Austin-Cohen formula which seems to hold well for daylight transmission over sea water thru a fairly wide frequency range. A simple calculation discloses that at a distance of 100 kilometers (62 miles) from a 750-kilocycle (400-meter) transmitting station, the sea water factor (0.0015) gives a received field strength about seventy times as great as that given by the dry sand factor (0.028). This strikingly illustrates the wide variation in the attenuations caused by different surface conditions.

City Transmission

In cities, where large steel buildings and similar structures enter as factors in transmission, the situation is much the same as over open country except that the effects are more intense and of smaller dimensions. For instance, the fact that heavy shadows can be cast is well illustrated by the curves in Figure 5.

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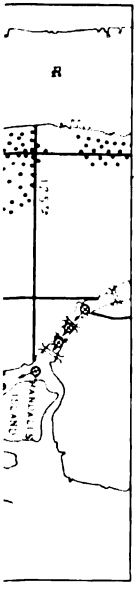

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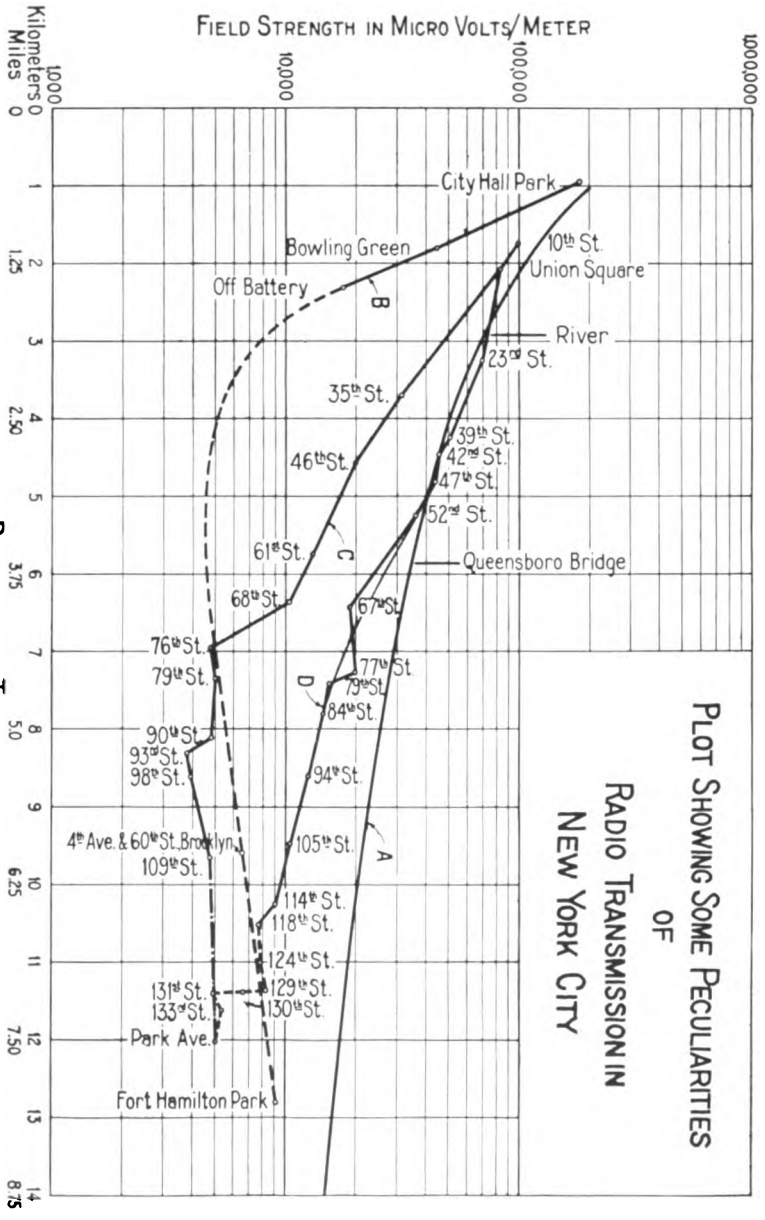

FIGURE 5—Plot showing Some Peculiarities of Radio Transmission in New York

sweep. At their feet the earth sometimes tends to drag them down, and sometimes casts obstacles in their way, but the gaps are quickly healed as the greater store of energy above is partially directed downward

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The way in which this self-healing action occurs is brought forward strikingly in the case of transmission northward along Manhattan Island. In Figure 5, Curve *C* gives measurements taken on a line running northward along Park Avenue. The first point at Union Square is just in front of a group of fairly large buildings, the effect of which is registered in a sharp dropping of the field strength. Beyond this, running up the center of Park Avenue, the drop is more uniform and less rapid. The high attenuation of the buildings along the avenue effectively prevents any appreciable healing up of the initial drop. At about 68th Street, a group of big apartment buildings and the crest of a slight hill cause another irregularity. The rest of this curve can best be considered in conjunction with Curve *D*, which is for a line of points up the East River. The river points were taken on the upper deck of a small steamer. The field strength holds up well until the Queensboro Bridge is approached. The low point, and the subsequent recovery to a smooth curve, account for the shadow caused by the bridge. The fact that the field strength over the river is much larger than that over the adjacent city causes a continual feeding in of energy from the river to the land. This is manifested in three ways:—(1) the recovery and flattening out of Curve *C* as the action becomes pronounced, (2) the fact that Curve *D* shows a higher attenuation than for clear over-water transmission, and (3) the direction of wave arrival on land, as indicated by the arrows on the map, is deflected away from the transmitting station and toward the river. The definite swinging round of the waves at the boundary line of two areas of different characteristics is closely analogous to the refraction of light. At its outer end Curve *D* is shown dotted because here the points depart radically from a radial line to the station and follow up the Harlem River to the place where the two lines cross and the curves come together.

Having thus seen with a considerable degree of clearness how the impediments to transmission operate in certain outstanding cases, and having a considerable number of measured points in and about New York and Washington, we have had the temerity to draw what may be called radio field-strength contour maps of the two cities. It must be admitted frankly that these maps are at best only an approximation, and that in many areas the data is too meagre to do more than vaguely support the contour lines. But a careful inspection will usually disclose a surprising agreement between the local territorial features and the values of the measured points, as to where the contour lines should lie.

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On the maps the measuring locations are indicated by crosses. Where loop direction was taken the cross is elongated into an arrow denoting the apparent direction of wave arrival. The value of field strength in milli-volts per meter is written beside the cross or arrow. The contour interval is not uniform thruout, but the value of each contour in milli-volts per meter is indicated in the usual way by a number inserted in a short gap in the line.

Figure 6 shows one of these maps embracing only the island of Manhattan. An outstanding characteristic is the relation of the contour lines to the Park Avenue and East River measurements discussed above. In the river, the contours bend rapidly inward toward the island until at some places on land in the northern part of the island the contours are actually parallel to the radial line from the transmitting station.

This does not mean that the wave fronts are necessarily traveling perpendicular to their starting direction; in fact, there is some doubt as to the meaning of the term "wave front" in such a case where there must be a conglomeration of reflected, refracted, diffracted, and re-radiated elements. The direction from which the energy is apparently arriving is, however, given by the loop position for maximum received current. The arrival directions, as was said above, favor the river. They also tend toward perpendicularity to the contour lines. In places where the wave travel is smooth and unobstructed, this becomes more strictly true—the wave fronts and the contours coincide, and the loop direction is perpendicular to both.

On the west side of the city, the same heavy attenuation on land and the same feeding in from the river is present as on the east side; in fact, it is even more pronounced at some points. At a few points, such for instance as at Columbus Circle, the loop direction does not follow the general trend. In most of these cases there was evidence of heavy local distortion of the waves due to the presence of a nearby building which was resonant at the transmitting frequency used.

As might be anticipated from the foregoing discussion, Central Park, surrounded by the city on all sides, has a low area or dead spot. Several contours close on themselves within this area and the field strength at one point drops to a value as low as it has 30 miles (48 km.) or more out in the country.

Another low point is indicated near Spuyten Duyvil. This one lies down in a valley behind a hill, and is probably due to the shadow of the hill.

In order to bring out more clearly how the various types and

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heights of buildings affect the transmission, we have plotted the contours on an aerial photograph of New York City. This is reproduced as Figure 7. Because of the location of the airplane when the photograph was taken, Manhattan Island is very much fore-shortened. In the original plot, which is made on a large photographic print, individual buildings may easily be seen, so that the reasons for many of the peculiarities of the contour lines are evident.

Figure 8 gives a radio field strength contour map of the whole Metropolitan area. From this map, the high radio attenuation of the thickly settled city is unmistakable. It gives the contours a roughly dumb-bell shape. Several other features are notable. To the west, a marked indentation of the two outside contours shows the effect of the city of Newark. This is substantiated by only a few points, but the points were carefully checked. To the northeast, the contours loop out over Long Island Sound indicating low attenuation over the water. Perhaps the most interesting feature of this map is the way in which the 10- and 20-milli-volt contours approach and recede from each other. Starting at the north on Manhattan, where they are closest, and proceeding clockwise, they spread gradually until in the residential district of lower Brooklyn they are well apart. Then, the abrupt indentation of the inner one and the slight indentation of the outer one show the extent of the shadow of lower Manhattan. For the rest of the circuit, except for the effect of the city of Newark, they spread widely apart in response to good transmission over the New Jersey meadows.

New York City, from the transmission and distribution standpoint, presents a "horrible example" of the difficulties which may be encountered. On the other hand, the situation at Washington, District of Columbia, is a fair example of an excellent distribution. The complete contour map of the Washington area is given in Figure 9. While the effect of land and water is well marked on this map, the deviation of the contours from the circular form which they would have in the ideal case is not sufficient to constitute serious irregularity. At only one point, to the west of Washington, is there an indication of a real dead spot. This is substantiated by only a single point. The measurement at this point was carefully made, but it was not checked and no other measurements were taken nearby. We were unwilling, therefore, to give the contours a radical shift to account for the single unsupported point and have left them undrawn thru this locality.

From the information which this study has disclosed many

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interesting and useful conclusions may be drawn. Some of them are as follows:

1. The radio attenuation over different kinds of earth surface varies widely. It is low for sea water and for flat moist ground. For dry ground, the attenuation is relatively much greater. In the case of closely built cities filled with steel buildings the local attenuation may be enormous.

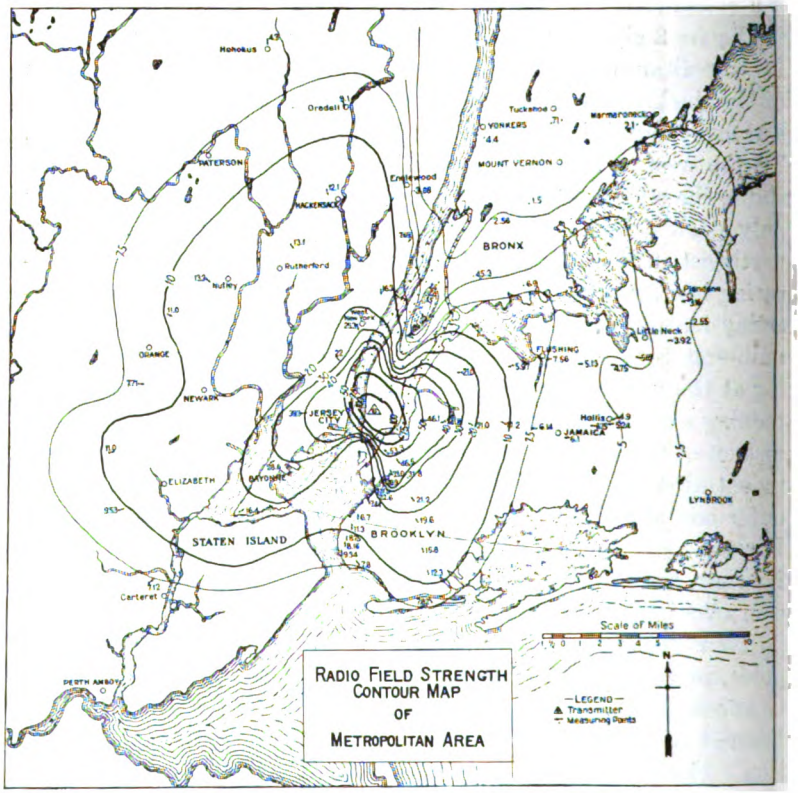

FIGURE 8—Radio Field Strength Contour Map of Metropolitan Area
(Field Strengths in Millivolts per Meter)

2. Sudden changes in land elevation and large masses of conducting material cast radio shadows which may be very heavy in extreme cases.

3. Shadows cause local dead spots; but usually within a relatively short distance beyond, the shadow is wiped out by refraction or diffraction.

All of these effects can be predicted by purely theoretical considerations, so it is not astonishing to find them. It is impos-

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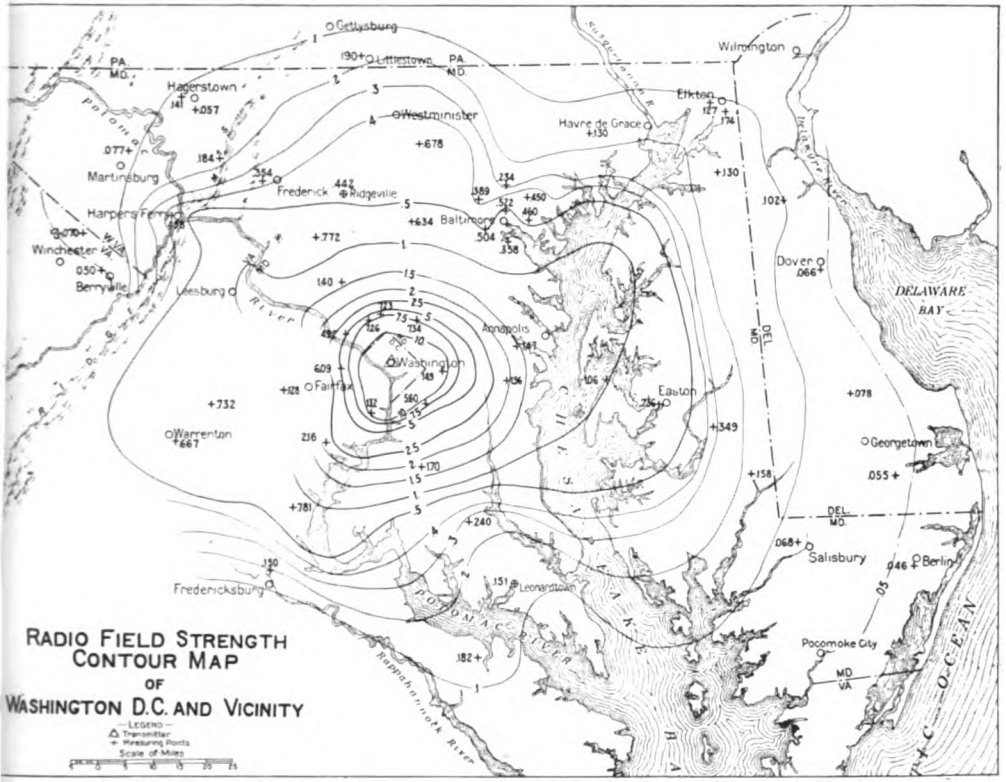

FIGURE 9—Radio Field Strength Contour Map of Washington, District of Columbia and vicinity

sible to predict the magnitude of the effects, however, on account of the irregular and complex nature of the transmission media and the lack of quantitative data on their electrical characteristics. The experimental study not only shows the magnitudes for specific cases but, as a whole, throws the picture into focus and provides the mind with a sort of scale or measuring stick which is useful in estimating the probable effects in other similar situations.

SUMMARY: **The paper presents the results of a quantitative study of radio distribution in and about the cities of New York and Washington, District of Columbia. The wide variations of attenuation over different kinds of territory, the causation of radio shadows or dead spots, and the phenomena of refraction and diffraction are all illustrated by the data and curves. Radio field strength contour maps of the two territories are presented.**

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THE MARCONI FOUR-ELECTRODE TUBE AND ITS CIRCUIT*

By

H. de A. Donisthorpe

(New York Representative, Marconi International Marine Communication Company, Limited)

Before I commence the subject of my paper this evening, I should first like to express my pleasure at having the honor of reading a paper before this Institute of world-wide reputation, the journal of which is read with such avidity by all the leading men in the science of radio.

I think that I am correct in saying that up to the present time little or nothing has been done commercially on this side of the Atlantic with the four-electrode tube. However, I understand they are being experimented with at the present moment by the General Electric Company. In Great Britain these tubes have found considerable favor amongst the radio engineers, and this type of tube is now a current store issue to all the large vessels of the British mercantile marine where the radio installations are under the care of the Marconi Company.

In order that this paper may be as complete as possible, it will perhaps be of interest to trace the evolution of the modern quadrode and to describe some of the early types of this form of tube which are interesting if only from a historic point of view.

The classification of the different types of quadrodes now in existence is not an easy matter, as each individual valve may be said to have been evolved for some particular purpose; that is to say, some are utilized for the purpose of rectification only, and others for magnification of radio signals, while others are designed for the reception of continuous waves.

Roughly these tubes may be classified into three separate groups, this classification being only useful from a constructional point of view. In every case, however, two of the electrodes function as controlling elements.

This construction classification can be detailed as follows:

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Type	Types of Electrodes
Group 1	Tubes with Two Grids in addition to the usual filament and anode
Group 2	Tubes with Two Anodes in addition to the usual filament and grid
Group 3	Tubes with Three Anodes in addition to the usual filament

Due to the presence of this additional electrode in the various types of quadrodes, the function of the electronic stream varies considerably from that of the well-known thermionic current produced between the hot cathode or filament and the anode of the popular triode. It is for this reason that the quadrodes present an interesting study. In some types of four-electrode tubes there is a diversion of electrons from one main electrode to another; the electrons during this period do the useful work required to produce the resulting effects, whilst in another instance the electrons are projected by virtue of their high velocity to a destination more distant than the usual anode of the triode.

The origin of the quadrode can, I think, be safely attributed to the genius of Majorana, who took out a patent in the United Kingdom for his tube in the year 1912. The circuit in which this tube was employed and the general arrangement of the four electrodes within the tube are shown in Figure 1. The tube in this instance functions purely as a rectifier, the quality of which is rather poor.

It will be seen from the figure that the grid itself consists of two separated and insulated portions, *A* and *B*, which form two separate grids in addition to the usual filament and anode and, therefore, comes under Group 1 of the above table. These two grids, it will be noticed, are interposed between the remaining two electrodes. The circuit in which the oscillations to be rectified are taking place is shunted across these two elements *A* and *B*.

A potential difference is then established between *A* and *B*, due to the oscillations taking place in the circuit, the result of which is to bring about an attraction and repulsion alternatively of electrons to and from *A* and *B*. The electrons are, therefore, diverted from their true path, namely, towards the anode with

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a resulting diminution of the filament anode current. This brings about variations of current in the external telephone circuit which is directly connected with the anode and a partially rectified current of the induced radio oscillations is effected.

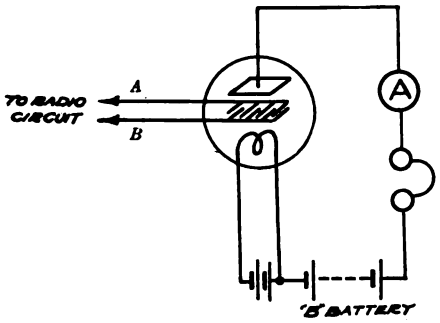

FIGURE 1—Majorana's Quadrode and Circuit

In view of the fact that this is a potential-operated device it is necessary to employ a large inductance across *A* and *B* in order to produce maximum efficiency, and consequently this particular type of tube is most sensitive when employed for the reception of long waves.

Another tube which functions in a very similar manner to that of Majorana but of a somewhat later date is that due to Professor Fleming. This particular quadrode comes under Group 3, and consists of three anodes in addition to the usual hot cathode or filament.

The mechanical construction and attendant circuit is shown in Figure 2. The tube, it will be seen, consists of four metal plates which are curved along their axes so that their convex faces are presented towards the filament, the filament being a short, straight vertical wire made of tungsten situated in the middle of the four plates.

The one pair of diametrically opposite plates *A* and *B* are joined together externally and constitute one single electrode and are known as the collecting plates. The second pair of plates *C* and *D* are connected directly across the receiving circuit and are the controlling elements functioning similarly to the two grids in the Majorana tube.

The function of the tube can be briefly described as follows:

The incoming radio oscillations in the receiver produce potential differences across *C* and *D* and thus diminish the established filament-anode current towards *A* and *B*, bringing about the

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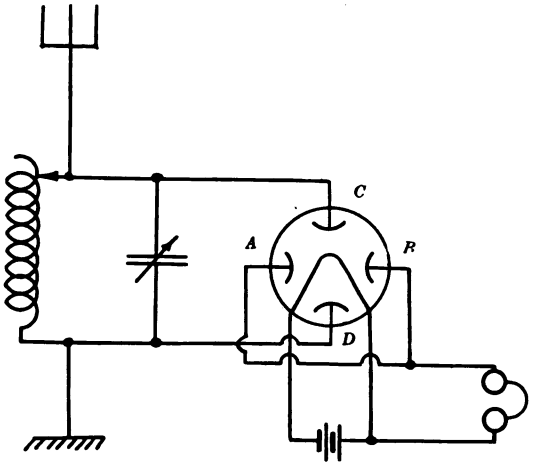

FIGURE 2—The Fleming Quadrode and Circuit

same results as those achieved within the Majorana tube. The characteristic curve illustrating this section is shown in Figure 3, which shows clearly that the decrease in filament-anode current

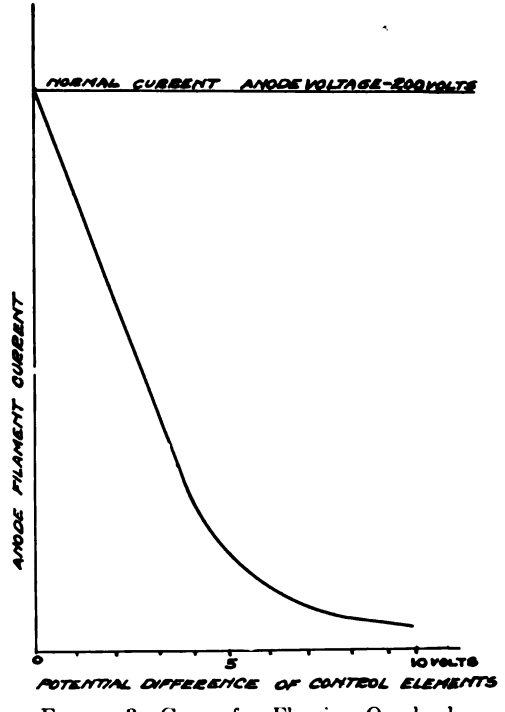

FIGURE 3—Curve for Fleming Quadrode

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is determined by the potential difference across plates *C* and *D*, so that the results obtained are proportional to the amplitude of the radio oscillations being received.

Practically speaking the only difference between the Fleming and Majorana tubes is in their construction and position of electronic disturbance.

I now come to the main feature of my paper in the description of the Marconi quadrode or FE I type of tube as it is commercially described. This instrument is of particular interest since it is now a commercial issue to radio installations for shipboard use.

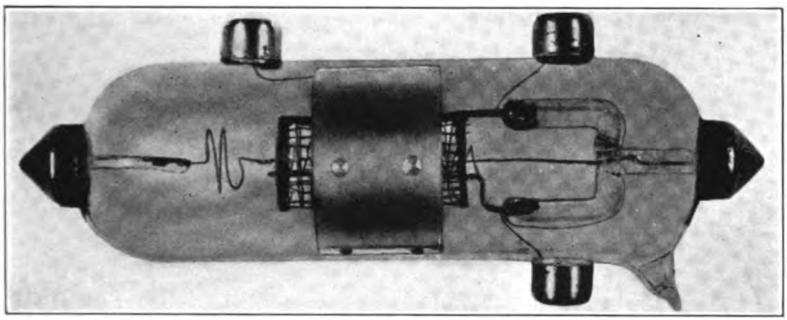

FIGURE 4—The Marconi FE I Quadrode

This tube is used in conjunction with a special amplifier known as Amplifier type number 91, Figure 5, and effects a one-stage radio frequency amplification, rectification, and a one-stage audio-frequency amplification, using the single tube, and consequently this quadrode may be said to do the work of three triodes.

Figure 4 illustrates the construction of the tube, the dimensions of which are about 3¾ inches (9.5 cm.) in length and ⅞ inches (2.2 cm.) in diameter. The following table gives the electrical constants of this tube.

Voltage Across Filament	Working Filament Current	Maximum 2nd Grid Current	Anode Voltage
4.5	1.2 amp.	0.75 milliamps.	45 volts

The tube, it will be seen, contains in addition to the usual grid (which is of a spiral nature) an additional grid which is of a

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wide open mesh form and, therefore, comes under Group 1 of the classification table. The electrodes in this tube are known as the "first grid," "second grid," and "outer electrode," and external connections are made to these electrodes by means of metal pips secured to the outside of the glass by means of plaster of Paris. Excluding the outer electrode this tube produces the usual characteristic curve of a fairly straight line, the second grid operating as the anode of a triode.

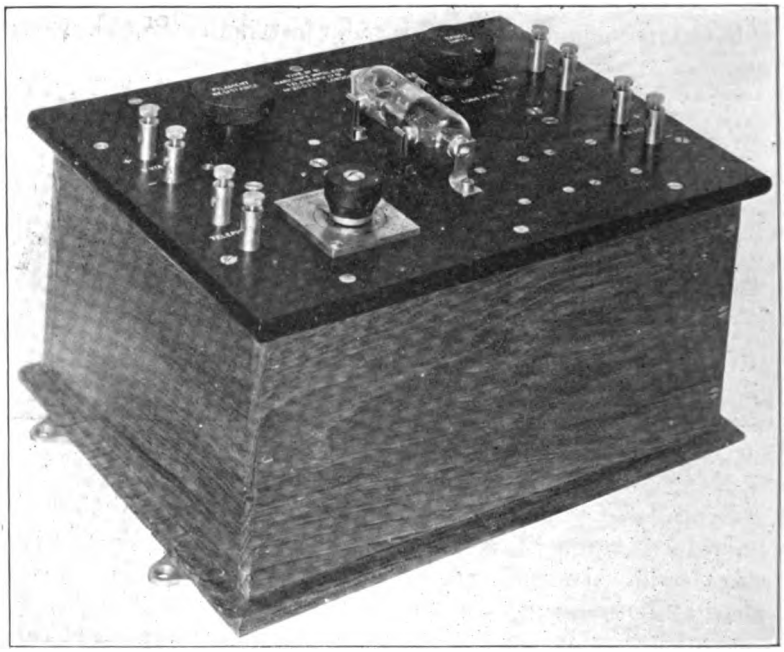

FIGURE 5—The Marconi Amplifier Type 91 for use with Four-Electrode Tubes

Having described the tube, it is next essential to trace the circuits of the amplifier in order to illustrate clearly the functioning of this particular type of tube. This type of amplifier will be of particular interest at the present moment when the interference to broadcasting is being so freely discussed by the press and Government Departments dealing with radio, for the simple reason that this amplifier, whilst producing a high degree of magnification, is not associated with any irradiation effect.

Figure 6 illustrates the circuit when the tube is employed purely for rectification purposes. It will be noticed that the second grid circuit corresponds to the usual anode circuit of a triode arrangement and includes the usual high tension or "B"

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battery for the purpose of establishing a steady filament-anode current. Now, if the first grid be given a positive charge with regard to the negative limb of the filament, the electrons will travel towards the second grid with a high velocity, and having due regard to the constructional design of this electrode, the electrons, because of their inertia, will pass thru it.

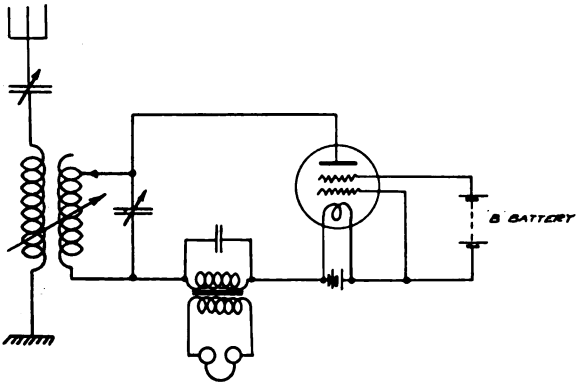

FIGURE 6—Rectification Circuit for FE 1 Quadrode

Now, the outer electrode, it will be seen, is connected to the negative limb of the filament and is, therefore, of the same potential, so that immediately an electron passes thru the second grid, it comes under the retarding electrostatic charge of this outer electrode.

Since there is a drop in voltage between the outer electrode and the second grid equal in value to the rise in voltage between the filament and first grid, the speed of the electron on its arrival at the outer electrode will be equal to the speed of the electron at the filament.

Assuming the initial speed of the electron at the moment of emission from the filament to be zero, the velocity of the electron on arrival at the outer electrode will also be zero. Now, if the resulting alternating emf., due to the induced radio oscillations, be impressed between the outer electrode and negative limb of the filament, a current will flow in that circuit whenever the outer electrode is made positive by those halves of the oscillations giving that potential, as when this condition is effected, the electrons are naturally attracted towards the outer electrode.

This function, it will be seen, is nothing more or less than a Fleming two-electrode tube rectification and the results are comparable to those obtained with a crystal. The rectification cur-

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rent arriving at the outer electrode has a saturation value of about 80 micro-amperes. However, with the use of the 91 type of amplifier, the oscillations are first submitted to a one-stage radio frequency amplification before the above form of rectification is effected.

The circuit embodying these two functions utilizing the one tube is shown in Figure 7. It has already been pointed out that this second grid of this type of quadrode functions according to the principle of the anode of a triode. It is, therefore, possible

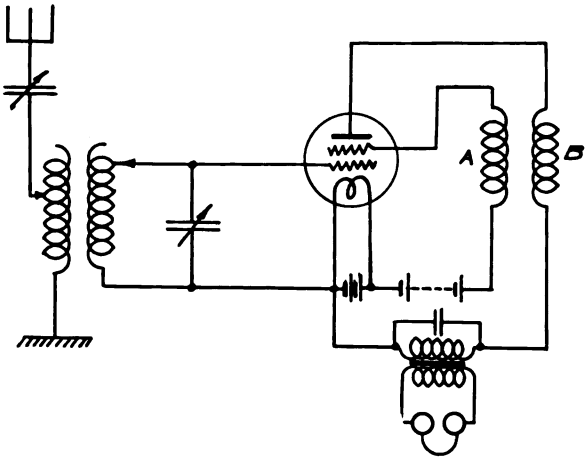

FIGURE 7—Radio Frequency and Rectification Circuits for FE 1 Quadrode

to produce large variations of filament anode current in the second grid circuit for small variations in potential of the first grid. This condition naturally occurs when the potential on the first grid is adjusted so that the tube is working on the straight part of the characteristic curve, and the magnified current produced in the second grid circuit will have the same shape and will be of the same frequency as that of the induced radio oscillations. In this circuit there is included the primary *A* of a radio frequency transformer, the secondary *B* of which is in the outer electrode circuit. It follows, therefore, that the magnified radio oscillations taking place in the second grid circuit will produce an alternating emf. across the secondary of the radio frequency transformer, or, in other words, across the outer electrode and negative limb of the filament; and consequently rectification is satisfactorily effected after the manner described above.

It is possible to submit these rectified and magnified signals

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to a one-stage audio-frequency amplification. This is obtained by impressing the rectified signals back on the first grid by means of an iron core transformer as is shown in Figure 8. The windings of this iron core transformer have the ratio of 1-to-1, and it will readily be seen that a simple low frequency triode arrangement follows in the first and second grid circuits. The audio-frequency oscillations are produced in the second grid circuit in the usual manner, in which circuit is located a telephone transformer. The windings, which are located in the circuits where radio oscillations are taking place, are naturally shunted with by-pass condensers.

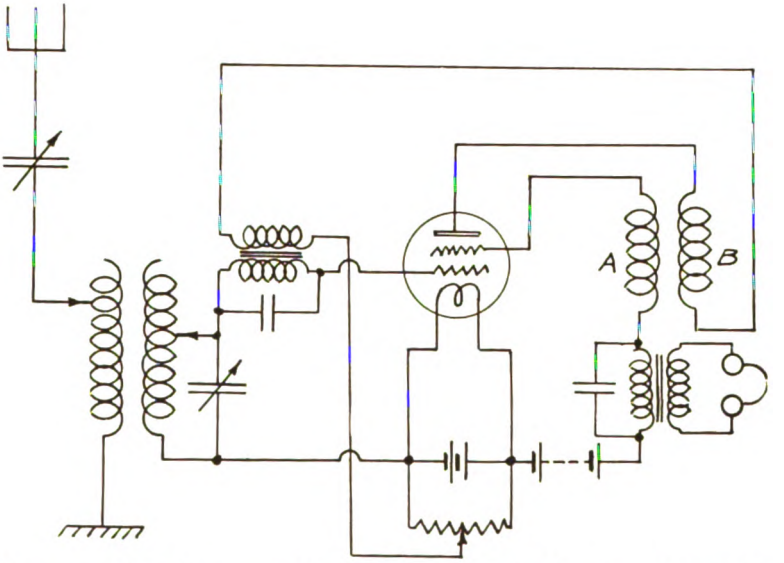

FIGURE 8—Radio Frequency, Rectification and Audio Frequency Circuits for FE 1 Quadrode

A feature of this type of amplifier is an adjustment, within limits, for the elimination of "jamming" or interference. For this particular purpose a potentiometer is employed, the high resistance winding of which is connected directly across the filament or "A" battery and the slider to the outer electrode thru the various elements of that circuit. This allows of alterations in the potential of the outer electrode being effected relative to the filament, and the well-known "limiting" effect can be satisfactorily produced.

The actual instrument is designed to cover a range of wave lengths from 200 meters to 2,800 meters on spark signals and up

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to about 12,000 meters on continuous waves. Above this wave length, the radio frequency amplification falls away slowly to about 2-to-1 at 20,000 meters.

Figure 9 shows the actual connections of the amplifier. The high frequency transformer *T* 1 has triple windings, all in series, each set having an equal number of turns in the primary and secondary. The first set has 220 turns each, the second 330, and the third 1,625. In the "short wave" position the switch short-circuits the second and third sets, in the "600" position the third set is short-circuited, and in the "long wave" position all are in circuit.

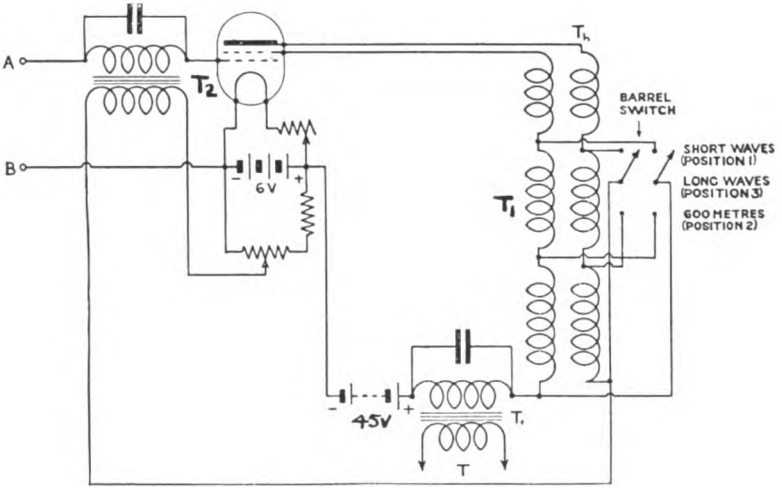

FIGURE 9—Connections of Type Number 1 Amplifier (Marconi)

Each winding of the transformer *T*, which is the closed iron core type, consists of 3,000 turns.

It will be noticed that the slider or contact of the potentiometer used for the "anti-jamming" adjustments operates over only half the total resistance, which is 220 ohms in all. This is to prevent this particular adjustment from being too critical. It will be found necessary, when using this adjustment, to vary at the same time the filament current within small limits.

The actual life of the tube is found to be in practice about 1,300 hours, altho, as a general rule, it is in excess of this, and one of these tubes has a record of 10,000 hours.

The amplifier itself has been criticized by some as not giving a high rate of magnification, but in this connection it is necessary to point out that the instrument gives a maximum increase of

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signals for feeble radio oscillations and does not respond so readily to strong signals, an ideal condition for shipboard reception. It is also useful from an economic point of view in that it renders maximum amplification with a minimum number of tubes, thereby reducing maintenance charges.

This instrument, which is fitted on a fleet of vessels running between England and Australia, regularly receives signals from the Leafield Station whilst these vessels are in Australian waters.

I think that this example alone will conclusively prove the efficiency of this type of tube as a commercial asset in the radio world.

SUMMARY: The paper deals firstly with the early forms of quadrodes or four-electrode tubes and leads up to a complete description of the Marconi type of quadrode or F. E. I. type and the attendant amplifier circuit.

The amplifier described utilizes one F. E. I. tube and effects a one-stage radio frequency, a rectification, and a one-stage audio frequency.

Information is detailed as to the practical experience obtained with this amplifier and quadrode, this particular combination now being a commercial issue to ships the radio installations of which are controlled by the Marconi International Marine Communication Company.

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THE PERFORMANCE AND THEORY OF LOUD SPEAKER HORNS*

By

Alfred N. Goldsmith, Ph.D.

Fellow A. I. E. E., Fellow I. R. E., Chief Broadcast Engineer

and

John P. Minton, Ph.D.

Fellow American Physical Society, Fellow I. R. E. Acoustic Engineer

(Technical and Test Department, Radio Corporation of America, New York)

The loud speaker is an essential part of a radio receiving set. The poor performance of either will result in unsatisfactory reception. The theory and operation of the latter have been dealt with in numerous places but the acoustic problems of reception, even tho of great importance, have received comparatively little attention. We believe, therefore, the present article dealing with the loud speaker is one which will prove of considerable interest to radio engineers.

A few years ago, Professor G. W. Stewart[1] published data on the performance of conical horns used as acoustic receiving apparatus. It will also be of interest to publish similar data on conical horns used for transmitting or producing sounds rather than for receiving. The present paper contains data which are of practical importance since they cover the range of horn lengths and solid angles which are apt to be considered in practice.

The curves show sound pressure produced at various frequencies. The ear is primarily a pressure device, and for this reason the pressure units of sound can be associated directly with the loudness as perceived by the brain.

Method of Test

The method of testing consisted in placing the horns in a well-padded sound proof booth and measuring the sound pressure at a point on the axis of the horns about 15 cm. (6 inches) from the mouth of the horn. A measuring system consisting of a con-

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[1] "Physical Review," pages 313-326, 1920

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denser transmitter, a five-stage resistance-coupled amplifier feeding into a thermocouple and galvanometer was calibrated to enable sound pressure to be determined from the galvanometer readings. Tests were made at several different frequencies to ascertain the manner in which the pressure along the axis varied with the distance from the opening of the horns. These test-curves showed that the sound pressure varied inversely as this distance except within a few centimeters of the opening and within a few centimeters of the walls of the booth. From about 10 to 75 cm. (4 to 30 inches) away from the mouth of the horns the end effect and the reflection from the walls were of no consequence. The pressure then decreased with the distance just as it would have done in the free, unconfined space. The curves shown in Figure 1 are given to illustrate these results. The curves show that a point 15 cm. (6 inches) from the opening is satisfactory to avoid room effects. The dotted lines are calculated values.

All the tests were made with the same receiver—namely, one of high impedance—operated at a constant voltage of about 15 volts thru 20,000 ohms. The results, therefore, are strictly comparable. A three-electrode tube oscillator was used to supply the current to the receiver attached to the horns. The oscillograms are included to show that the voltage and current waves are practically pure sine waves. The oscillograms also show the sound output from the horns as picked up by a condenser transmitter and amplified by a six-stage resistance-coupled amplifier. Some of the curves were taken at resonant frequencies of the horns and others were taken at anti-resonant frequencies. Hence, in the former case, any impurities are masked by the intense output at resonance whereas in the second case the harmonics show up prominently. No trouble from impurities occurred above, say, 400 cycles. Our curves, then, at low frequencies are sufficiently correct at the resonant frequencies, but are not so at the low anti-resonant frequencies. At low frequencies, then, we must take these factors into account. Reference to these oscillograms will be made later.

Horns

Sixteen conical horns were made of heavy galvanized iron. Sufficient rigidity of the walls was secured to avoid material vibrations. The results can be taken as closely representative of horns with rigid walls and, therefore, the data are characteristic of the air columns themselves.

The following table summarizes the horn-dimensions in cm.:

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Axial length	31	34	34	34	66.3	65.3	65 0	63.0	124	123	122	120	183	182	180	180
Diam., small end	1 6	1 6	1 6	1 6	1 6	1 6	1.6	1 6	1.6	1 6	1 6	1 6	1 6	1 6	1 6	1.6
Diam., large end	5 0	10 3	15.3	20 3	7 6	15.7	30 5	45 5	15.5	31.0	45	60	15	30	60	90

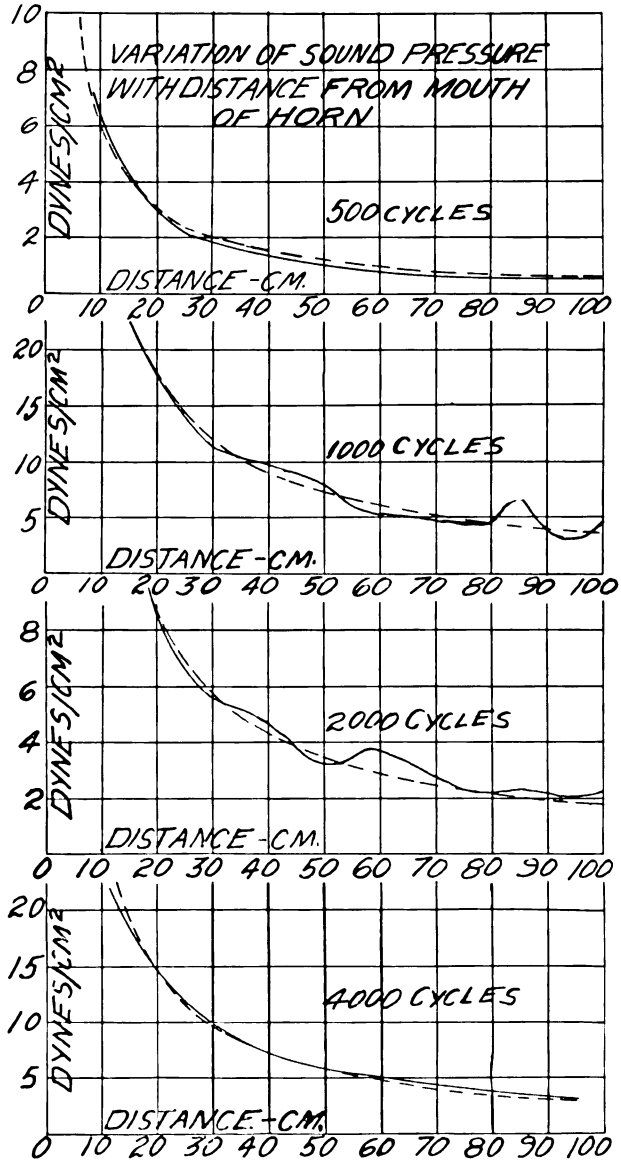
VARIATION OF SOUND PRESSURE WITH DISTANCE FROM MOUTH OF HORN
500 CYCLES
1000 CYCLES
2000 CYCLES
4000 CYCLES
DYNES/CM²
DISTANCE - CM.
0 10 20 30 40 50 60 70 80 90 100

FIGURE 1

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Experimental Results

The curves given in Figures 3, 4, 5, and 6 for the sixteen horns are shown in four groups; each group contains the curves for horns of constant length with varying opening. A comparison can also be made with constant opening and varying length. Still a third comparison can be made with constant solid angle at various lengths. The comparisons can be made at constant frequencies or at the fundamentals and also at the various overtones of the horns.

In Figure 2 is given the sound output of the receiver unit without a horn attached to it. The condenser transmitter was placed 15 cm. (6 inches) from the opening of the unit to the surrounding space. For this reason we can gain some information

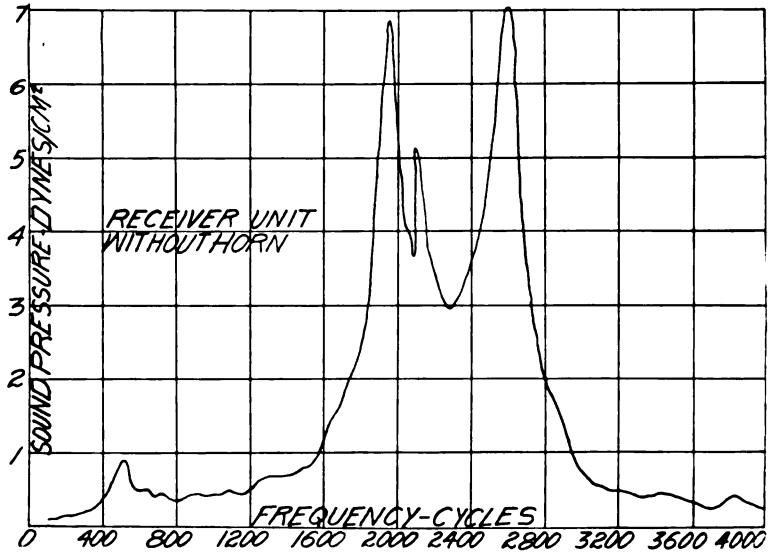

Figure 2

as to the amplification produced by the horn. We can also use this receiver unit curve to assist in the interpretation of the curves. The fundamental of the receiver diaphragm as shown by the curve is 500 cycles. The peaks at the higher frequencies are produced by the diaphragm vibrating in some of its higher modes of vibration. These effects will be modified some by the various horns and our curves give a comparative basis for the horns with the unit used.

The first group of curves gives the pressure-frequency curves for the horns of the shortest length, namely 34 cm. approximately.

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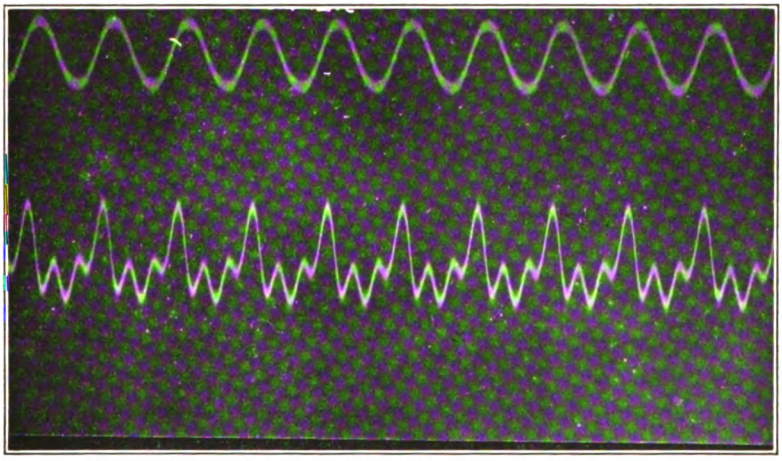
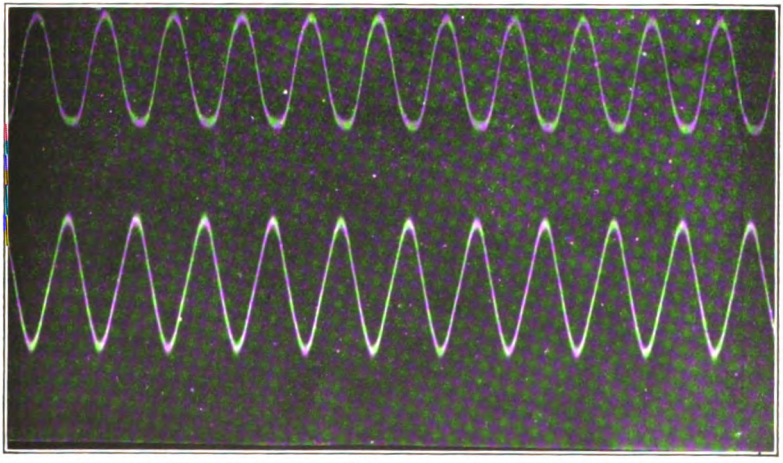
These are shown in Figure 3. Several characteristics are distinctly noticeable. There are pronounced peaks in all the horns. The two large ones in the neighborhood of 500 and 2,300 cycles are due to the receiver diaphragm; the others (except some of the "ripples" below 300 cycles) are due to the resonating properties of the horns. The fundamental on the basis of its wave length equals four times the length of the horn and should occur at about 270 cycles. The overtones—the odd ones only being present in a horn closed at one end—should occur at 810, 1,350, 1,890, 2,430, 2,970, 3,510, 4,050 cycles, and so on. The first horn peak, however, is not 270, but 320 cycles. The two oscillo-



Current Thru Unit—Sound Output—Plate I (1), 269 Cycles



Current Thru Unit—Sound Output—Plate I (2), 320 Cycles

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grams, shown in Plate I, show that the 269-cycle peak is composed almost entirely of the double frequency term. This is due to the second harmonic introduced by the unit itself, which corresponds to the natural frequency of the diaphragm. The oscillogram for the 320-cycle peak indicates that this peak is composed almost entirely of the fundamental.

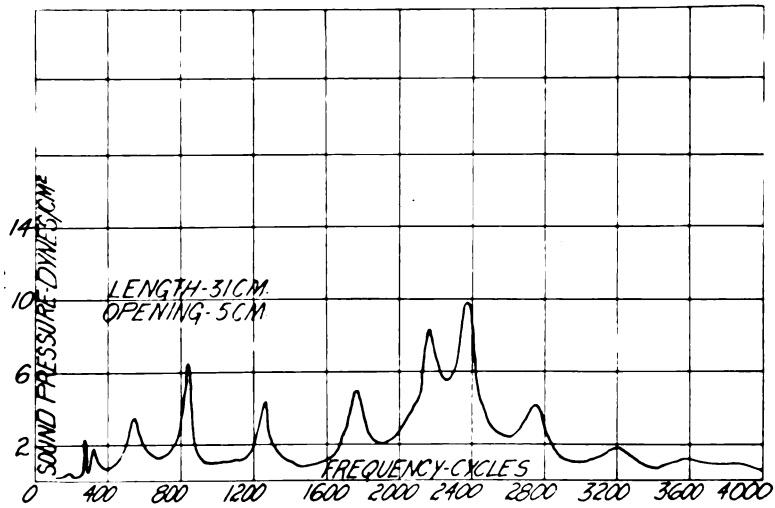

FIGURE 3—GROUP I

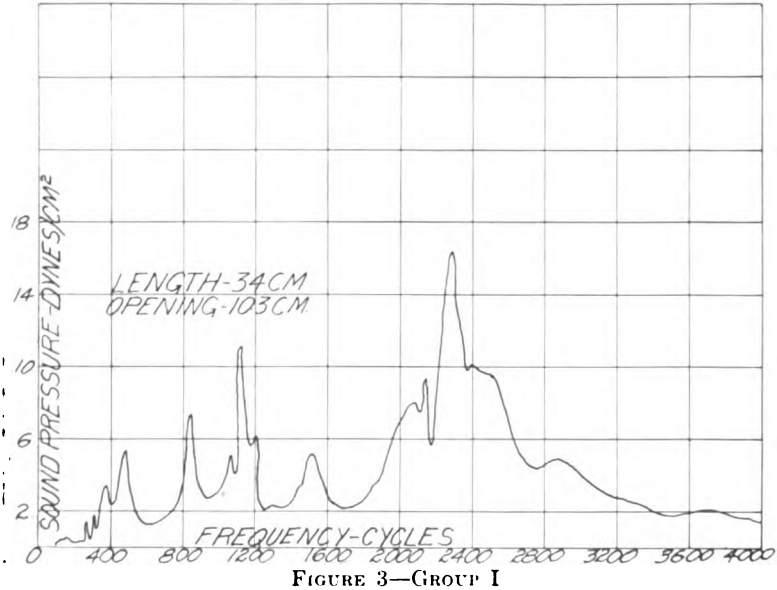

FIGURE 3—GROUP I

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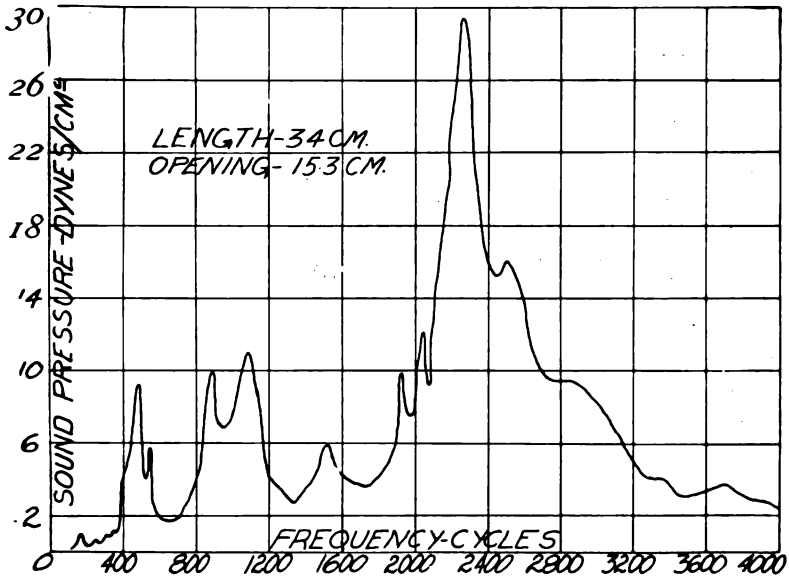

FIGURE 3—GROUP I

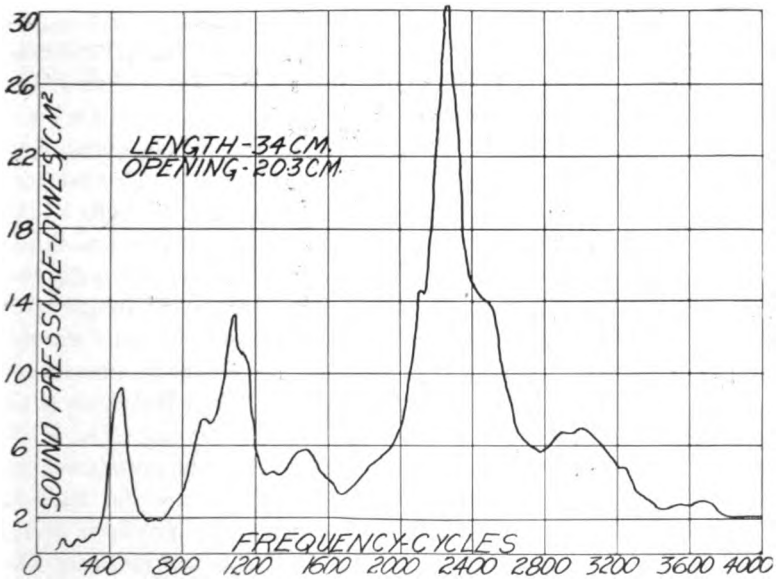

FIGURE 3—GROUP I

Reflection of sound at the smaller openings is considerable, and marked resonance will occur with these openings. Reflection is much less for the larger openings, and, hence resonance

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will be less marked in these cases. Furthermore, as pointed out by Professor Stewart,[2] reflection takes place along the entire inner surface of the horn and therefore "resonance" will occur at all frequencies above the fundamental.

This phenomenon is most pronounced for horns with large openings, the length being kept constant. Two effects may therefore be anticipated for horns of large openings. Firstly, the peaks at definite resonance frequencies in the curves should be less prominent because there is less end reflection in the horns with large openings. And secondly, the large "valleys" in the curves should be relatively filled in or elevated because there is increased reflection along the walls of the horns, and therefore, more reinforcement of *all* frequencies above the fundamental.

With this viewpoint as a guide, it is interesting to study the peaks. They are sharp and pronounced with the horn of the smallest opening and less sharp and pronounced with the large opening. Using an end correction of 0.7 times the radius of the opening as recommended by Stewart, we find that for the smallest horn there should be 8 horn peaks from 253 cycles, the fundamental, to 3,800 cycles, the 15th overtone (if we call the fundamental 1). Assuming a simple closed pipe the peaks should occur at the following frequencies: 253, 760, 1,265, 1,770, 2,270, 2,790, 3,290, and 3,800. Of course, the resonant peaks will follow a closed pipe only roughly. The agreement is not exact, but it is satisfactory. What has been said about the smallest horn may be repeated in substance for the next larger size. Here, however, the end correction is larger, and the peaks should occur at 221, 665, 1,110, 1,545, 1,990, 2,430, 2,870, 3,310, 3,760 cycles, and so on. Since the end reflection vanishes for the higher frequencies with the larger openings, there should be no marked resonant frequencies in this region. About all that is actually observed is a "wave," so to speak. Such waves are, however, for example, present at the last four frequencies, mentioned above. The 1,990-cycle peak is interfered with by its proximity to the receiver peak. The 1,100 and 1,500-cycle peaks are properly located, but the lower ones are not. The fundamental and first odd overtone are too high to check with the simple theory for pipes which we are here using. Apparently the end correction of 0.7 radius is too large for the low resonant frequencies. The 270-cycle peak is a second harmonic one—the first horn peak being 320 cycles.

Horn resonance is less marked for the 15 cm. opening and still less marked for the 20 cm. opening. As for the average sound

[2] I revious citation, page 324.

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output, however, the smallest horn gives the least, the next large size still more, and the two larger ones about equal outputs with perhaps a preference for the horn with the 20 cm. opening. One is not to conclude, however, that the horn with the least amount of resonance and largest opening will give the largest output. The data for the longer horns indicate an optimum opening for best results, in accordance with Professor Stewart's results.[3]

The basis for consideration of the longer horns is the same as that given for the shortest ones tested. It should be remembered in studying these curves that the sound outputs at the low frequencies are small. Overtones, therefore, which are always generated in the receiver are particularly disturbing for the low frequency measurements. A peak may exist at a low frequency reading but the sound pressure may be due to a considerable extent to the overtone. This effect is not important, however, above 300 cycles.

The data for the 65 cm. horns are plotted in Figure 4 in Group II. As we would expect, resonance is pronounced, the peaks are sharp, and the ratios of maxima to minima are large for the horns with the 7.6 cm. and 15.7 cm. openings. The horn resonance is much less marked for the 30.5 cm. opening and is practically absent for the 45.5 cm. opening. The average sound output from the largest aperture horn of this length is about the same as that from one having the smallest opening. The two intermediate horns give superior results, the larger of the two intermediate sizes have less resonance and a more uniform output. As in the case of the shortest horns, the ratio of maxima to minima approaches unity for the higher frequencies and for the largest openings.

The peaks for this group of horns should be about 200 to 250 cycles apart. The fundamental for these 65 cm. horns should be in the region of 120 or 130 cycles. The curves show small peaks in this region. However, the vibrational energy of the diaphragm is so small at low frequencies that a small sound output must nevertheless be the result. This is still true for the long horns tested. The horn will be of little value for low frequencies, even if it is a long one, unless the receiver unit itself has at least a reasonable amount of vibrational energy at those frequencies. This shows the importance of having a low natural frequency unit to give low frequencies. On the other hand, the low frequency unit will not vibrate with any intensity, at the high frequencies, which would rather indicate the desirability of a high frequency

[3] I revious citation.

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unit. Since, for high frequencies, reflection will occur at the end of the horn if the opening is small, it means that the opening will need be rather small to take advantage of any resonance effects, when high frequency response is needed. In the theoretical part

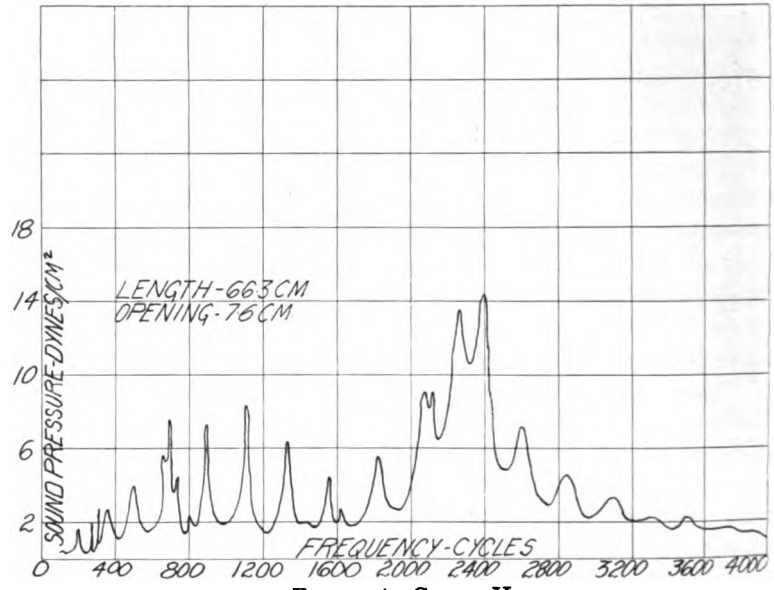

FIGURE 4—GROUP II

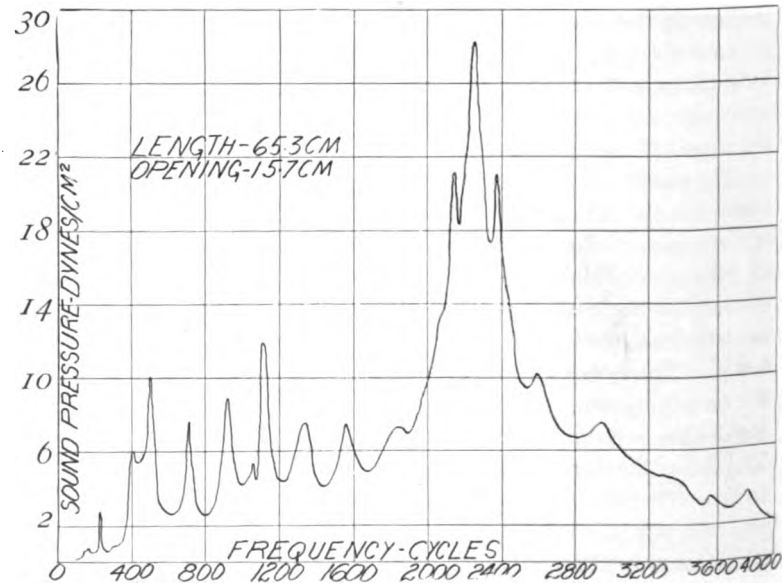

FIGURE 4—GROUP II

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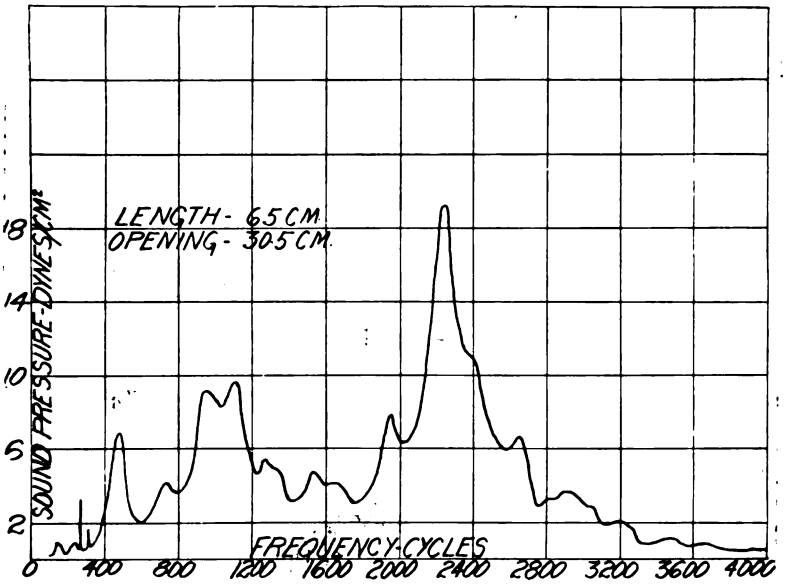

FIGURE 4—GROUP II

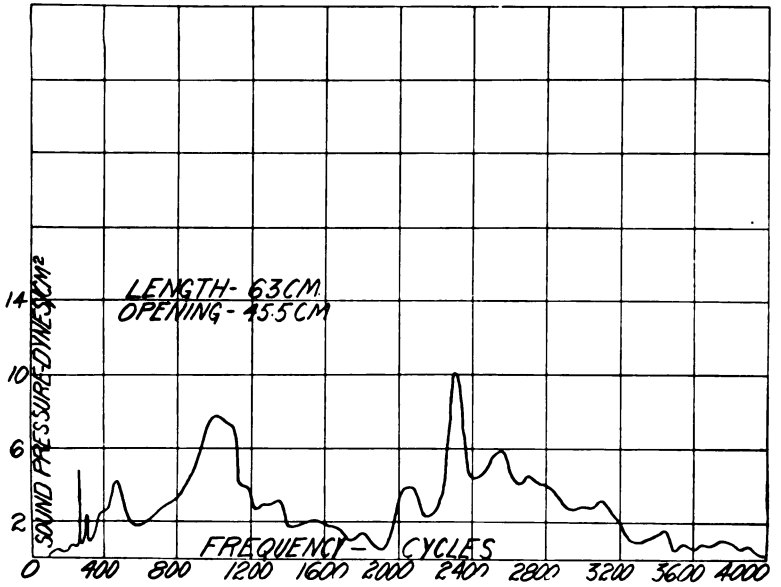

FIGURE 4—GROUP II

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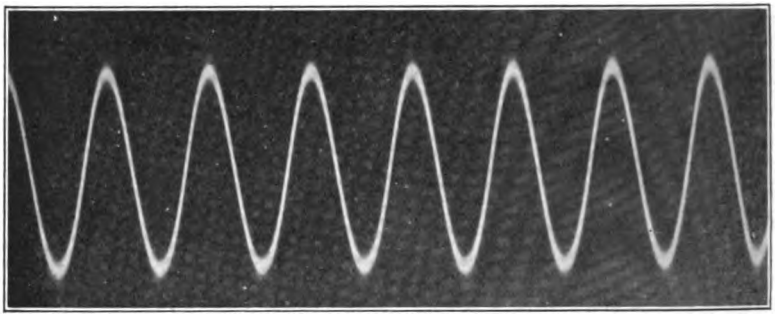
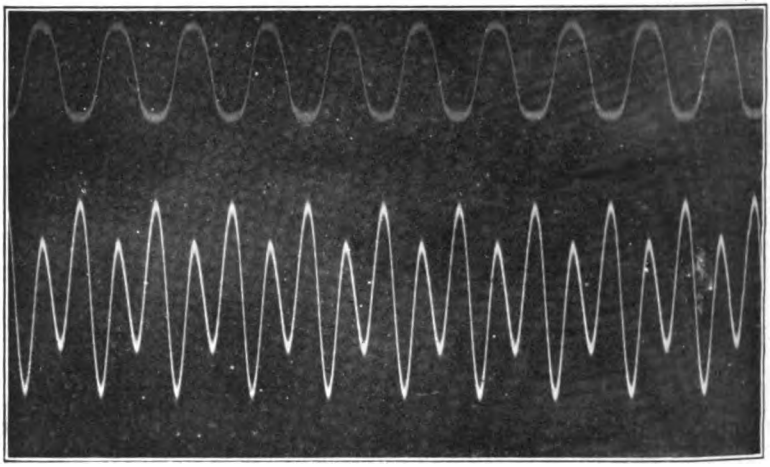
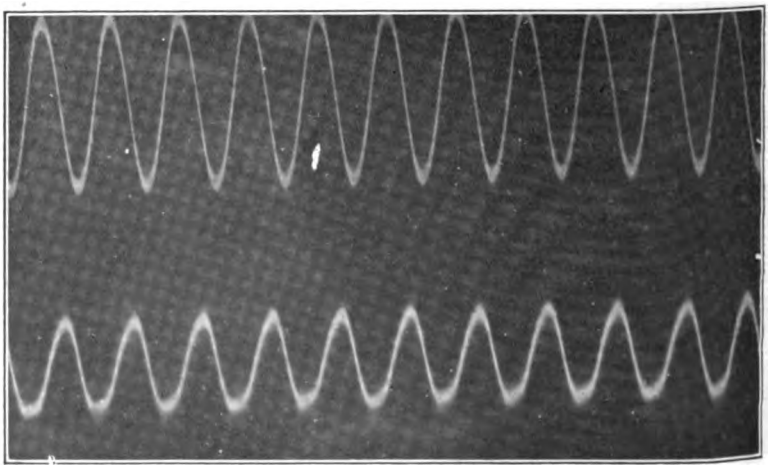


Sound Output—Plate II (1), 200 Cycles



Current Thru Unit—Sound Output—Plate II (2) 270 Cycles



Current Thru Unit—Sound Output—Plate II (3) 320 Cycles

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of the paper it is pointed out that the air velocity and pressure are largely out of phase at the low frequencies. This prevents large sound radiation at such frequencies.

In these 65 cm. horns we may call attention to the sharp peaks at 270 cycles. The oscillograms in Plate II show that these peaks are chiefly second harmonic. Those peaks at 200 cycles and 320 cycles, however, are chiefly fundamentals as shown by the reproduction.

An analysis for the 270-cycle oscillograms was made by Mr. E. W. Smith of this laboratory. The first of the two following equations is an analysis of the oscillogram itself, and the second equation is the analysis of the sound itself. The latter was obtained from the former by correcting for the frequency characteristics of the measuring system.

OSCILLOGRAM EQUATION

$$y = 1.00\ sin\ (\omega t + 301°) + 5.45\ sin\ (2\ \omega t + 338°) + 0.562\ sin\ (3\ \omega t + 77°) + 0.993\ sin\ (4\ \omega t + 325°) + 0.342\ sin\ (5\ \omega t + 222°) + 0.03\ sin\ (6\ \omega t + 45°) + 0.127\ sin\ (7\ \omega t + 105°) + 0.333\ sin\ (8\ \omega t + 320°) + 0.475\ sin\ (9\ \omega t + 153°) + 0.206\ sin\ (10\ \omega t + 248°) + 0.255\ sin\ (11\ \omega t + 65°)$$

and for the corrected sound output:

$$y = 1.00\ sin\ (\omega t + 301°) + 6.52\ sin\ (2\ \omega t + 338°) + 0.787\ sin\ (3\ \omega t + 77°) + 1.438\ sin\ (4\ \omega t + 325°) + 0.513\ sin\ (5\ \omega t + 222°) + 0.046\ sin\ (6\ \omega t + 45°) + 0.196\ sin\ (7\ \omega t + 105°) + 0.516\ sin\ (8\ \omega t + 320°) + 0.74\ sin\ (9\ \omega t + 153°) + 0.323\ sin\ (10\ \omega t + 248°) + 0.4\ sin\ (11\ \omega t + 65°).$$

As stated above, the 270-cycle peak is due chiefly to the second harmonic, nevertheless the analysis shows that other harmonics, even up to the eleventh, are present to a noticeable degree. These equations, then, serve to illustrate the significance of the distortion introduced by a receiver unit and horn and how the original sounds may be modified greatly at the receiving end.

The data for the 125 cm. horns are plotted in Figure 5 in Group III. The resonant peaks should be about 130 cycles apart, the fundamental being about 65 cycles on the basis of a closed end at the receiver. In the case of the horns of this length the peaks should occur roughly at 65, 190, 320, 450, 580, 710, 840, 970, 1,100, etc., where the horn is considered closed at the receiver end. The agreement of the actual positions of the peaks with the predicted positions is fair.

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For the Group III horns, the sound output was greatest for the 31 cm. opening just as was the case for the 65 cm. horns. The horn with a 15 cm. opening gave about the same average

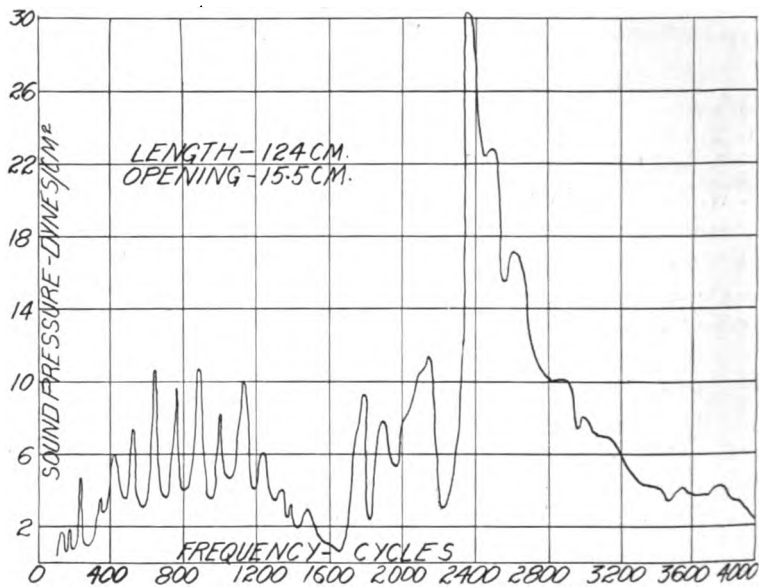

FIGURE 5—GROUP III

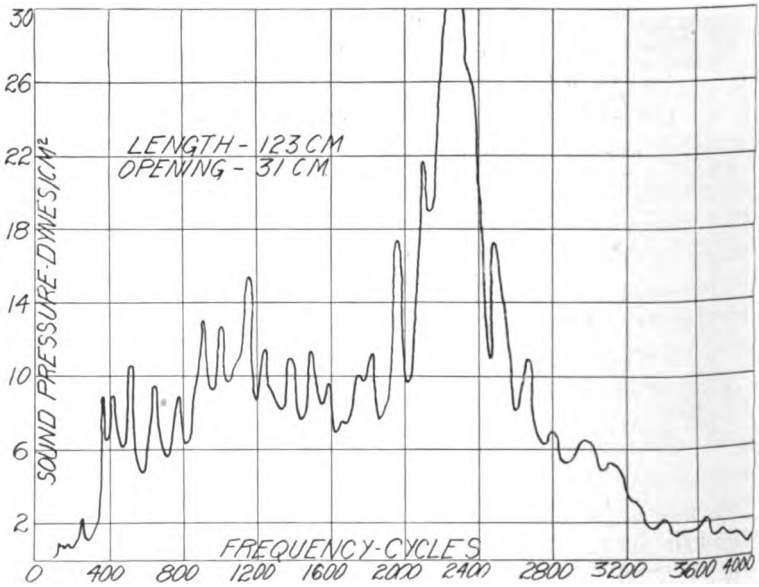

FIGURE 5—GROUP III

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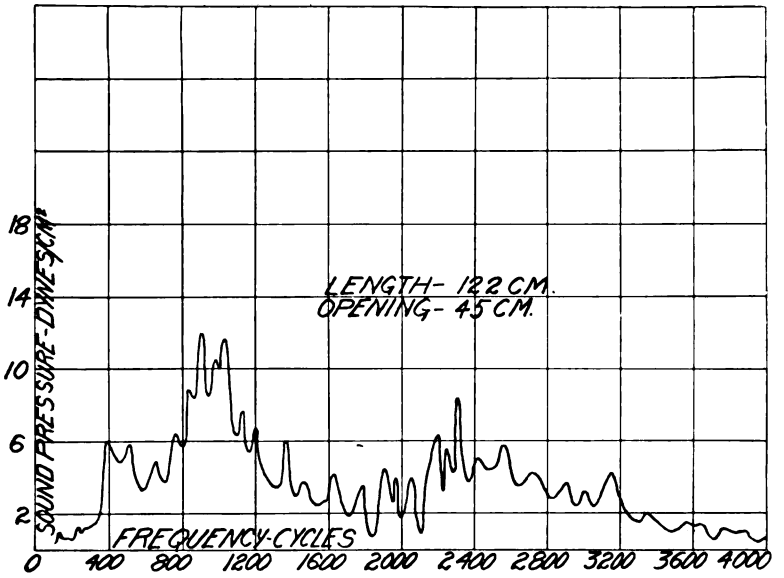

FIGURE 5—GROUP III

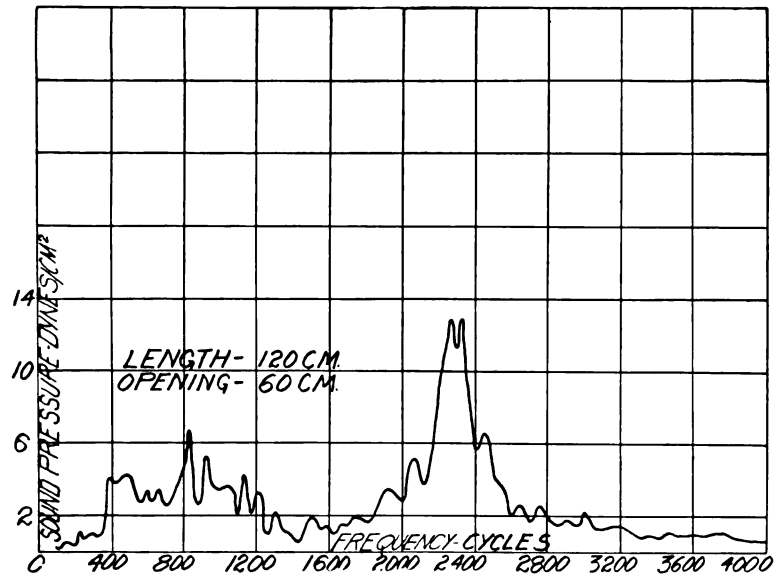

FIGURE 5—GROUP III

response as was obtained with the 45 cm. opening. The horn with the 60 cm. opening is quite inefficient compared with these of the smaller openings.

The data for the 180 cm. horns are plotted in the curves

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shown in Figure 6 in Group IV. The sound output from the horns with the 60 cm. and 90 cm. openings is small, the latter

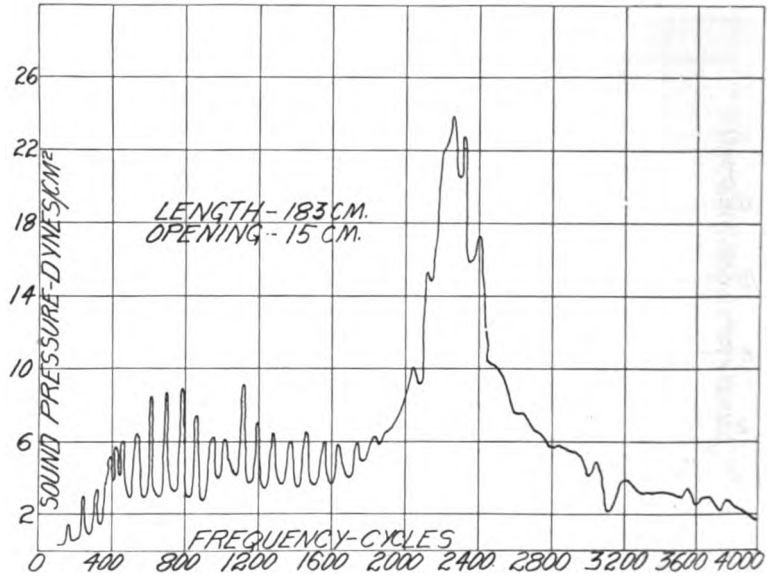

FIGURE 6—GROUP IV

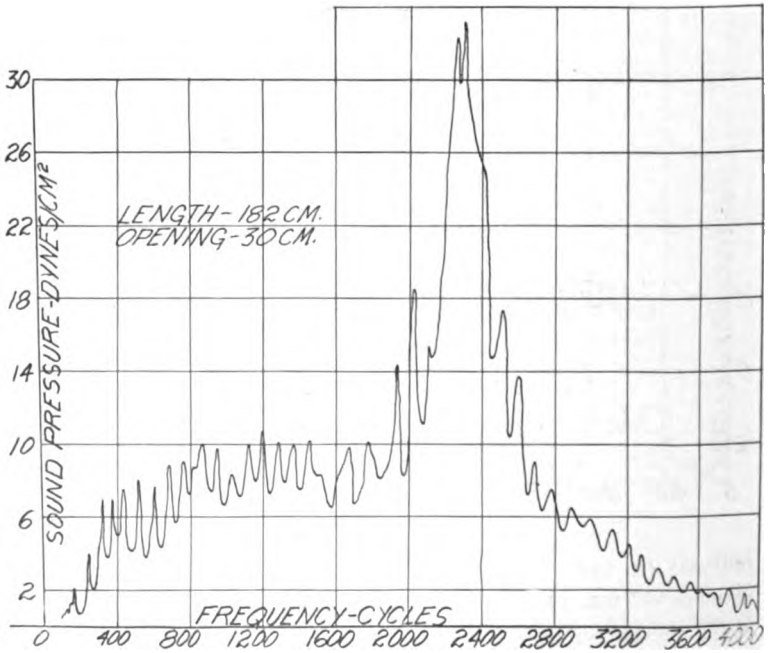

FIGURE 6—GROUP IV

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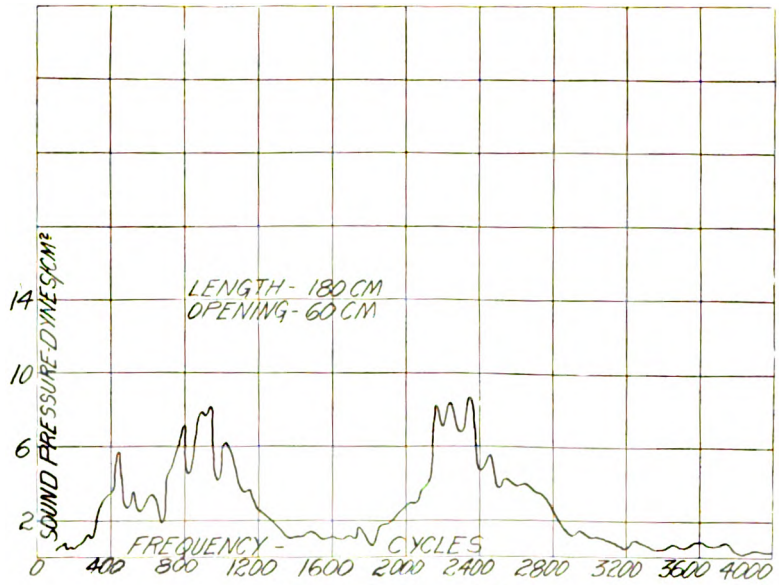

FIGURE 6—GROUP IV

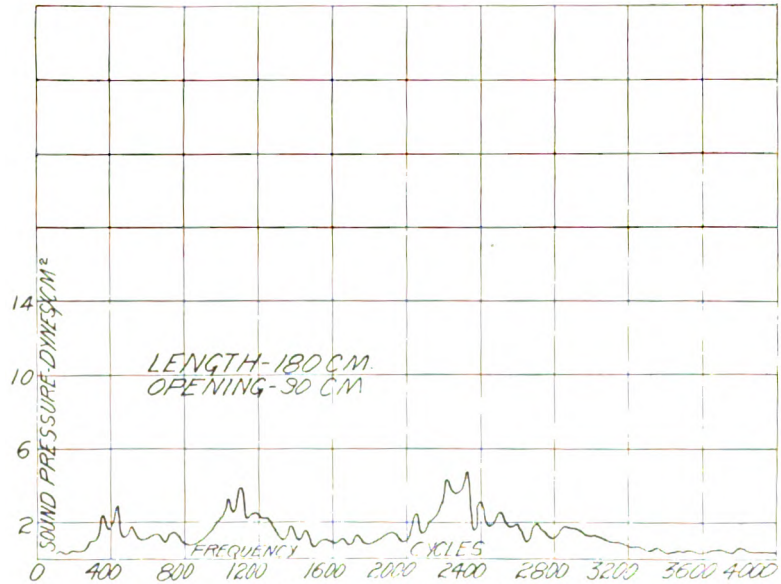

FIGURE 6—GROUP IV

horn giving the smaller output of the two. The horn with the 30 cm. opening gives much larger outputs than do those with the larger openings. The horn also gives larger outputs than does the one with the 15 cm. opening.

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There are pronounced resonant peaks for the two smaller size horns, but not for the horns with the 60 cm. and 90 cm. openings. On the basis of simple, straight pipe theory, the peaks should be about 80 or 90 cycles apart. This is in agreement with the separation of the peaks as shown in the curves. If the end is considered closed the peaks should occur at approximately 45, 135, 225, 315, 405, 495, 585, 675, 765, 855, 945, 1,035, 1,125, 1,215, 1,305, 1,395, 1,485, 1,575, 1,665, 1,755, 1,845, 1,935, etc. The actual location of the peaks is in quite satisfactory agreement with positions predicted on the basis of a pipe with one end closed. In discussing the peaks it is interesting and important, theoretically, to observe that the peaks for the various horns of constant length occur at the same frequencies. It is also important to observe that resonance plays a sufficiently prominent part to cause as much as 3-to-1 or 4-to-1 pressure variation in the output over a very small frequency range; and for the horns with the smaller openings the variations are sometimes as large as 8 or 10-to-1. Of course, such horns with sharp resonant peaks, due to large end reflections, are of perhaps quite limited value as far as a loud speaker horn is concerned.

We will next consider a comparison for constant opening and varying length of horn. Taking first the frequencies above 1,800 cycles, the curves for the horns with 15 cm. openings indicate that there is little preference among the four horns. The sound output is essentially the same for these conical horns having a 1.6 cm. initial opening and a 15 cm. final opening. The phenomenon of resonance plays no important, or at least no controlling part, and all of them have little or no damping effect on the receiver diaphragm for this frequency range. The output over this region for these particular horns is almost entirely a property, apparently, of the receiver itself.

Below 1,800 cycles the horn length plays an important part. The average response from about 400 cycles to 1,800 is about the same for the four horns with the 15 cm. opening, but the variation from this average is quite different in the four cases. The peaks are close together for the 180 cm. horn and far apart for the 35 cm. horn. The results over this region for the 180 cm. horn can perhaps be considered the most satisfactory of the four horns. The 125 cm. and 180 cm. horns are certainly more satisfactory than the two shorter ones over this low frequency region. This, of course, should be the case for the longer horns, simply because, by means of resonance, the low frequencies are amplified more with these longer horns than with the shorter ones.

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Also, as pointed out in the theory, for a given low frequency, the longer horn will have a better power factor at the large end than the shorter one, and for this reason, too, the longer horn will radiate more sound. The question of radiation and the effect of the matching of acoustical impedance will be considered in detail in this connection in the theoretical discussion.

A comparison of the three longest horns with a 30 cm. opening can be made. Even a casual comparison will exclude the 65 cm. horn as far as superiority of the average sound output is concerned. The superior performance of the 180 cm. horn is not only noticeable below 400 cycles, but is also noticeable at all other frequencies where the fluctuation from the average response, due to resonance for small frequency ranges (plotted as a function of the frequency) are relatively small. The cause of this is that the resonant peaks are close together. A horn of about this same opening but about 300 cm. long was constructed, and its behavior is given in Figure 7.

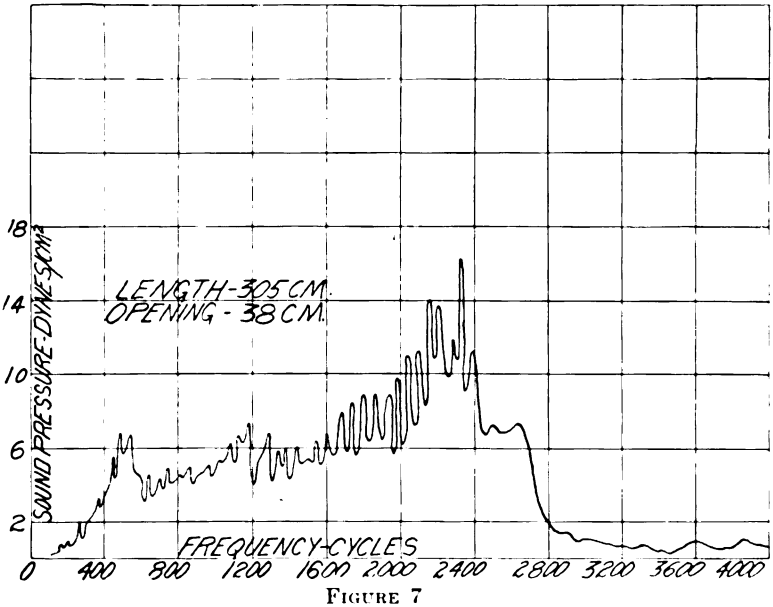

FIGURE 7

The curve for this long horn, which had a 38 cm. opening at the large end and a 1.6 cm. opening at the small end, shows resonant effects which are not very marked. The peaks are close together, being about 50 or 60 cycles apart as they should be for a horn of this length. Since the peaks are close together, there

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cannot be large depressions between the peaks. The range of response or width of each peak is sufficient to eliminate large fluctuations within small frequencies ranges. At high frequencies the output from this long horn does not compare favorably with that from some of the short horns. The output at the very low frequencies is no greater than that of some of the shorter horns either. As previously pointed out, a long horn cannot be expected to give a large response at low frequencies if the receiver unit itself is not designed particularly for low frequencies, even tho the power factor may be more favorable.

A comparison can be made between the two horns with 45 cm. openings and with lengths 65 and 125 cms. The latter is the more satisfactory, but neither of them gives a performance equal to that of either of the 125 cm. or 180 cm. horns with the 30 cm. openings. The effect of resonance, radiation, and acoustical impedance on these performances will also be considered in the theoretical discussion.

There is little to choose between the two longer horns with the 60 cm. openings. Neither of them is comparable in performance with the two longer horns with the 30 cm. openings.

As is to be expected the short horns with the small openings give unsatisfactory performance. We conclude, therefore, the optimum opening and length for the sixteen straight conical horns are 30 cm. and 180 cm., respectively. The practical importance of this conclusion is at once apparent.

REACTION OF HORN ON DIAPHRAGM

We do not desire to give in this paper an experimental and theoretical consideration of the reaction of the horn on the receiver diaphragm, nevertheless, it seems appropriate to give a set of curves for a typical impedance analysis. For this purpose, we include impedance tests for the receiver unit alone and for the unit when attached to the 125 cm. conical horn with a 31 cm. opening. These results are shown plotted in Group V. The tests cover the same frequency range as was covered in the tests previously described. In presenting these curves, perhaps we should call attention to some of the outstanding points. The curves of reactance and resistance are given—the impedance analysis of them are omitted. The curves show that the natural period of the diaphragm without the horn is about 500 cycles. When the horn is used the natural period decreases to 440 cycles. This change can be due either to an increase in the effective mass or a decrease in the effective elasticity of the diaphragm due to the horn effect. Perhaps there is no decrease in elasticity at all,

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but most probably an increase. Hence, the decrease in natural period is due entirely to an increase in the effective mass brought about by the added inertia due to the air confined in the horn neck.

We note that the horn peaks at 360 and 520 cycles affect the reactance and resistance of the receiver unit to an extent equivalent to about one-half the effect produced by the diaphragm resonance. That is, near the resonance of the diaphragm, where its motion is controlled largely by its mechanical resistance, the reaction of the horn produces reactance and resistance changes

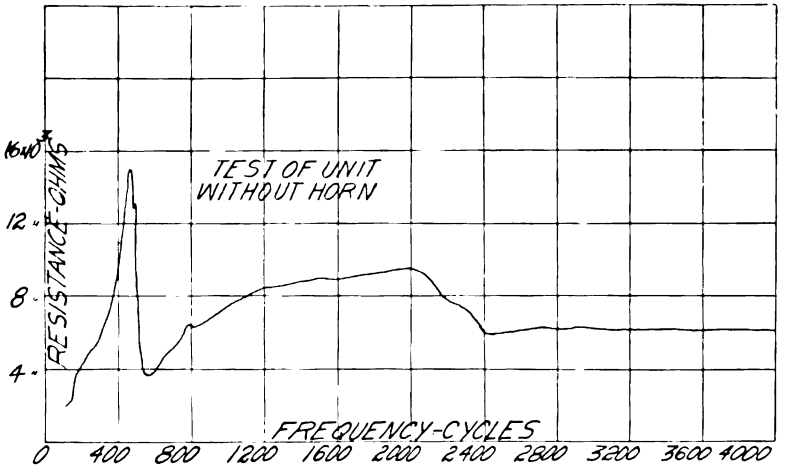

FIGURE 8—GROUP V

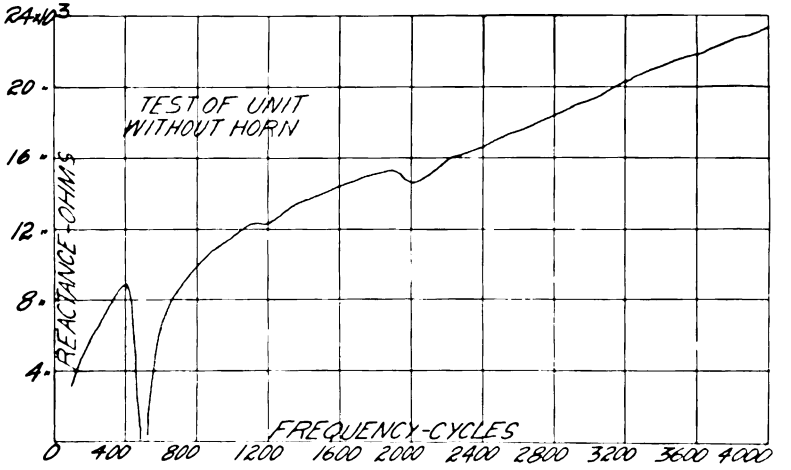

FIGURE 8—GROUP V

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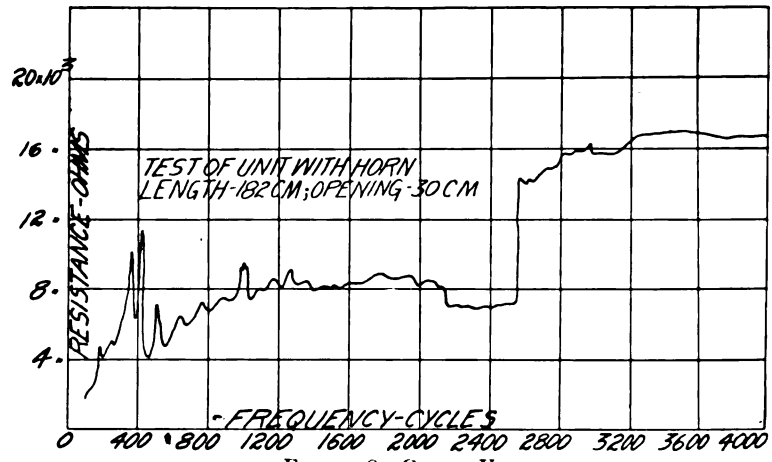

FIGURE 8—GROUP V

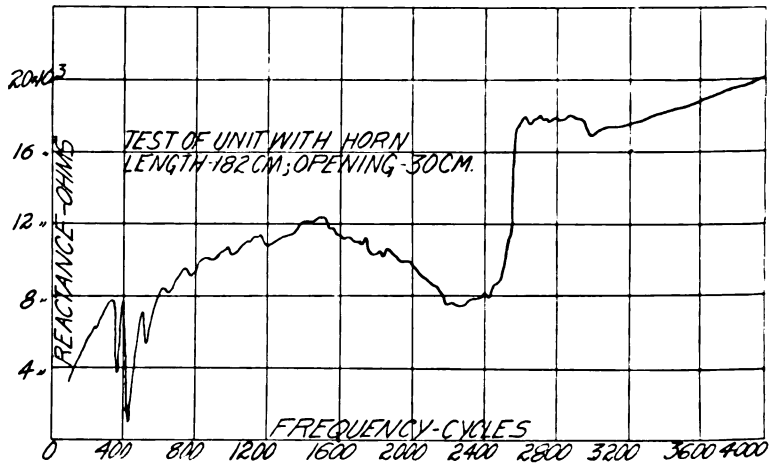

FIGURE 8—GROUP V

about half those produced by the diaphragm. The sound energy radiated, therefore, produces dissipation of energy about equal to that produced by the mechanical resistance of the diaphragm itself. Consequently we can study the horn effects by means of impedance analysis.

Away from the resonant frequency of the diaphragm, the motion is controlled largely by either the elastic or the inertia force and the resistance force plays a smaller part than near resonance. In the same way the horn reaction will be comparatively small away from the region of the diaphragm resonance. The group

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of curves shown in V brings out this point quite clearly. Nevertheless, the horn resonance has its effect, which is distinctly noticeable up to about 3,000 cycles. The marked change in the curves in the region of 2,500 cycles corresponds to the changes in the mode of vibration of the diaphragm. Perhaps these remarks are sufficient to illustrate the method of study employed and the magnitude of the effects brought about by the use of a horn. We shall hope to have more to say on this phase of the subject at a later date.

Experimental Conclusions

The following experimental conclusions are arrived at based on the results of the above data.

(1) Resonance of the air column plays a marked effect for all the horns tested.

(2) The frequencies at which resonant peaks occur are close to integral multiples of the fundamental frequency of the column.

(3) Most of the curves indicate that the fundamental frequency of a horn closed by a receiver corresponds to a horn length of one-quarter wave length. Peaks always occurred at the predicted frequency differences, as is to be expected, even up to high frequencies.

(4) The horns give considerable amplification for any frequency nearby equal to or greater than the fundamental. That is, the horns resonate for all frequencies equal to or greater than the fundamental due to reflections occurring in proper phase relations along the walls of the horns. The resonating effect is grestest for the end reflection.

(5) The greater the frequency, the more nearly does the ratio between the maximum responses and adjacent minima approach unity. That is, at the higher frequencies, there is little amplification due to the horn resonance and it is important to observe that the horns have little effect on the sound output in this region provided not too small an orifice and too long a neck are used at the receiver end.

(6) Desirable results were obtained with those horns having a 30 cm. opening. Among the four horns having this size opening the best results were obtained with a horn of 180 cm. length.

(7) Low frequencies can be increased first, by resonance, provided the horn is long enough to bring its fundamental resonant frequency into this region; second, provided the receiver diaphragm will respond to the low frequency currents, and third, by obtaining a better power factor between pressure and current with a long horn.

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(8) The horn cannot of itself introduce the sound of the various frequencies. These must be present in the receiver unit in order that the horn may produce an appreciable re-enforcement of them.

(9) A long horn of appropriate shape, orifice, and end opening promises marked amplification of sound without highly marked resonant peaks thruout a considerable frequency range.

(10) Horn reaction on receiver diaphragm due to confined air and to sound radiation is marked, particularly near diaphragm resonance. For the present, however, conclusions based on impedance analysis are deferred for future publication.

Theoretical Considerations

The theoretical basis for the performance of horns of finite lengths and various shapes has been given by Professor A. G. Webster.[4] Prof. G. W. Stewart[5] extended Webster's theory to include additional results on conical horns. The present writers present in this paper a further extension of Webster's work to include the exponential shape for finite lengths. A review of the straight pipe and cone of unlimited lengths is given, and some original work on unlimited exponential and parabolic shapes is included. This is followed by a discussion of horns of finite lengths.

We are interested in the amount of sound energy that can be radiated into the space surrounding the horn. If a pipe or horn is of infinite length so as to avoid any end-reflection, and, if we assume no reflection along the walls of the horn, it is not difficult to derive the expressions for the radiant acoustical energy. This has been done by Lord Rayleigh.[6] Dr. Slepian and Mr. Hanna[7] have given an engineering interpretation to this work and extended it to cover horns of exponential shapes of infinite length, that is, those in which no end reflections occur.

1. Theory of Conical Horn of Unlimited Length

If we follow Rayleigh, the waves may be considered spherical and originating from a point and progressing outwardly. Their velocity potential is represented, therefore, by:

$$\phi = -\frac{A}{4\pi r}\cos K(ct-r) \qquad (1)$$

[4] "Journal of the National Academy Sciences," pages 275-282, 1919.
[5] "Physical Review," pages 313-326, 1920
[6] Previous citation.
[7] "Journal of the American Institute of Electrical Engineers"; mid-winter Convention, Philadelphia, February, 1924.

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where A determines the strength of the source, r is the distance out and $K = \frac{2\pi}{\lambda}$; λ being the wave length.

If we consider the case of a cone, with the sound source at the vertex, (1) is replaced by

$$\phi = -\frac{A}{\omega r} \cos K(ct-r) \tag{2}$$

where ω is the solid angle of the cone. The sound radiation is denoted by the work done per second by the sound pressure at the source forcing a "current" of air out thru the horn. The sound pressure is given by $-\rho \frac{d\phi}{dt}$, ρ being the air density. The total "current" per second is given by $\omega r^2 \frac{d\phi}{dr}$. Using these relations with (2) we get:

$$\rho \frac{d\phi}{dt} = \frac{\rho A K c}{\omega r} \sin K(ct-r) \tag{3}$$

and

$$\omega r^2 \frac{d\phi}{dr} = A\{\cos K(ct-r) - Kr \sin K(ct-r)\} \tag{4}$$

A comparison of these two equations will show that one component of the current is in quadrature with the pressure and the product of this component and the current, therefore, does not represent sound radiation or "loading" as used by Dr. Slepian and Mr. Hanna. The other component of the current is in phase with the pressure and their product represents power output in the form of sound radiation. This represents the "loading" as defined and used by the authors to whom we have just referred.

The phase angle between the pressure and the current is seen to be lagging and represented in Figure 9. The power factor is given by the cosine of the phase angle θ. It is:

$$\cos \theta = \frac{Kr}{\sqrt{1+K^2r^2}} \tag{5}$$

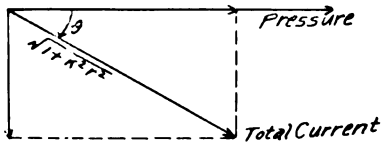

FIGURE 9

At frequencies sufficiently high or at values of r sufficiently large to make $K^2 r^2$ large compared to unity, $\cos \theta$ becomes unity

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as in the case of the tube. At low frequencies or at small values or r where $K^2 r^2$ can be neglected with respect to 1, $\cos \theta = K r$. It does not become zero therefore, until zero frequency is reached. We also note that when r is small, that is, near the source, $\cos \theta$ approaches zero. This does not mean, however, that the radiation approaches zero. On the contrary, it remains finite (see equation (6) below), because the pressure becomes increasingly large near the source to compensate for the change in power factor toward zero. This indicates, then, why it is difficult to obtain low frequencies with a horn.

The power output along the horn from the source of sound situated at the vertex of the cone, therefore, is given by the product of (3) and (4). The average value of sine-square being ½, the power output is:

$$P = \frac{\rho K^2 c A^2}{2\omega} \tag{6}$$

In the open space, of course, $\omega = 4\pi$. The power output, therefore, varies directly as the frequency squared and inversely as the solid angle.

2. Theory of Cylindrical Tube of Unlimited Length

We place the source of sound within the tube of cross sectional area σ. The energy is transmitted in both directions continually (that is, with no reflection) away from the source which we assume to be the same as that represented in equation (1). A short distance away from the source the sound waves are plane and are of the type:

$$\phi = A \cos K(c t - x) \tag{7}$$

Just as in the case of spherical radiation outward, near the source where r is small we have:

$$4\pi r^2 \frac{d\phi}{dr} = A \cos K(c t - r) \tag{8}$$

and also just as in the conical case where r is small (see equation 4)

$$\omega r^2 \frac{d\phi}{dr} = A \cos K(c t - r) \tag{9}$$

So, near the source situated in the pipe we have in the case of the unlimited pipe where all the energy is radiated through the total area 2σ

$$2\sigma \frac{d\phi}{dx} = A \cos K(c t - x) \tag{10}$$

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which represents the current passing along the pipe.

All three of the cases are similar, the area factor entering in some way for the sphere, the cone, and the pipe.

Whence,

$$d\phi = \frac{A}{2\sigma} \cos K(ct-x)\,dx \tag{11}$$

or

$$\phi = \frac{A}{2\sigma K} \sin K(ct-x) \tag{12}$$

Therefore

$$\rho \frac{d\phi}{dt} = \frac{\rho A c}{2\sigma} \cos K(ct-x) \tag{13}$$

and the pressure and current are in phase.

The power radiated in both directions from the source at $x=0$ is:

$$P = \left[\left(2\sigma \frac{d\phi}{dx}\right)\left(-\rho \frac{d\phi}{dt}\right)\right]_{X=0} = [\text{current} \times \text{pressure}] \tag{14}$$

This is similar to the relation in mechanics whereby the product of velocity, force, and time gives the amount of work.

Substituting (10) and (13) into (14) we get:

$$P = \frac{\rho A^2 c}{4\sigma} \tag{15}$$

where ½ has been substituted for $\cos^2 K ct$. As pointed out by Rayleigh, if a rigid barrier is placed in the tube near the source so that the energy is radiated in one direction only, the total radiation will be doubled, because by deflection the pressure itself is double while the total current remains unchanged. In this case

$$P = \frac{\rho A^2 c}{2\sigma} \tag{16}$$

The radiation, therefore, varies inversely as the cross section and is similar to the cone where we saw it varied inversely as the solid angle. That is, σ in the straight pipe case is analogous to the solid angle in the case of the cone. In the pipe, however, the radiation is independent of frequency and the current and pressure are in phase at all frequencies. Radiation, therefore, along an infinite pipe, is constant at all frequencies.

3. Theory of Exponential Horn of Unlimited Length*

In Appendix I we show how Professor Webster derived the equations for the exponential horn fron his general equations

*The parabolic horn is an interesting case and is treated in Appendix III.

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for horns of any profile whatever. In this appendix we show that, if ϕ is the velocity potential, $g=\frac{1}{2}\sqrt{4K^2-m^2}$, $K=2\pi$ divided by λ, the wave length, m is the exponential coefficient in the horn equation, area (σ) = initial area (σ_0) times ε^{mx}, then

$$\phi=\varepsilon^{-\frac{mx}{2}}[A\cos(nt-gx)+B\cos(nt+gx)] \qquad (17)$$

for the waves in both directions along the exponential horn. It is of considerable interest to note that the general solution for plane waves of the type

$$\frac{d^2\phi}{dt^2}=c\frac{d^2\phi}{dx^2} \qquad 18)$$

along straight tube in both directions as shown by Rayleigh and others is:

$$\phi=A\cos K(ct-x)+B\cos K(ct+x) \qquad (19)$$

Equations (17) and (19) are similar. $Kct=nt$, but Kx does not exactly equal gx for $g=\frac{1}{2}\sqrt{4K^2-m^2}$. If m^2 is negligible, or the frequency is high, then equations (17) and (19) are identical, except for the exponential factor in the solution of the exponential horn. This similarity between the exponential horn and the straight pipe was pointed out by Webster in his 1919 paper in the "Proceedings of the National Academy."

The solution of the exponential horn given in equation (17) is similar to the propagation of a wave motion in both directions along a string where a friction term is taken into account. This case is treated in Rayleigh's, "Theory of Sound," Volume I, page 232. The solution is also similar to the propagation of electricity in both directions along wires where the exponential factor in the present solution corresponds to the attenuation factor in the electrical case. Consequently, the exponential solution in the case of sound is due to the transverse vibrations which are not considered in the present approximate treatment.

If we assume that the total current flow into the small end of the exponential horn is the same, namely: $A\cos Kct$, as in the case of the straight tube and conical horn as already treated, then we may write for the total current at any place in the exponential horn (compare with (8), (9), (10), and (17)):

$$2\sigma_o\varepsilon^{\frac{mx}{2}}\frac{d\phi}{dx}=A\cos(Kct-gx) \qquad (20)$$

corresponding to:

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$$\phi = -\frac{A\,\varepsilon^{-\frac{m\,x}{2}}\,g}{2\sigma_o K^2}\left[-\sin(K\,c\,t-g\,x)+\frac{m}{2\,g}\cos(K\,c\,t-g\,x)\right] \quad (21)$$

Since the pressure is given by $-\frac{\rho\,d\,\phi}{d\,t}$ where ρ is the air density, we get from (21)

$$-\rho\frac{d\,\phi}{d\,t} = \frac{\rho\,A\,\varepsilon^{-\frac{m\,x}{2}}\,c\,g}{2\,\sigma_o K}\left[\cos(K\,c\,t-g\,x)-\frac{m}{2\,g}\sin(K\,c\,t-g\,x)\right] \quad (22)$$

In passing we desire to point out that these equations correspond to equations (78) and (79), published by Slepian and Hanna in their recent paper. The power radiation is given by the product of (20) and (22) and is

$$P = \frac{\rho\,A^2\,c\sqrt{K^2-\frac{m^2}{4}}}{2\,\sigma_o K\,\varepsilon^{\frac{m\,x}{2}}} \quad (23)$$

in which we have substituted the value of $g=\frac{1}{2}\sqrt{4\,K^2-m^2}$ In this product $\frac{1}{2}$ has been substituted for the square of the cosine and on the average the product of the sine and cosine terms does not represent power radiation.

A comparison of equations (20) and (22) shows that there is a leading current as shown in Figure 10 and by equation (24):

$$\cos\theta = \frac{1}{\sqrt{1+\frac{m^2}{4\,g^2}}} = \sqrt{1-\frac{m^2}{4\,K^2}} = \sqrt{1-\frac{m^2\,c^2}{4\,n^2}} \quad (24)$$

which is identical with equation (82) published by Slepian and Hanna. It is to be noted that the only characteristic of the exponential horn which is associated with the phase relation between velocity and pressure is that which determines the rate at which the horn opens up. The initial opening does not enter the phase relation. On the other hand, the initial opening of the exponential horn enters the pressure and power equation in exactly the same way the solid angle of the cone enters when the source of sound is situated at the vertex, and in exactly the same way the cross sectional area of the straight pipe enters Much care needs to be exercised, therefore, in making comparisons and drawing conclusions as to relative merits. Furthermore, our experimental data given in section I shows a distinct optimum opening and solid angle, and comparisons should properly be made between best results obtained with each of the various types. In the last analysis this simple theory does

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not check experimental data for the cone and we are quite certain that it will not check data on exponential horns. The theory has not gone far enough yet and we should be cautious about drawing conclusions from it. Nevertheless our theoretical deductions, therefore, are in agreement. We may emphasize some of the deductions that are to be drawn from this result. When $\sqrt{1-\frac{m^2}{4K^2}}$ is zero, the pressure and current are ninety degrees out of phase and, theoretically, no radiation results.

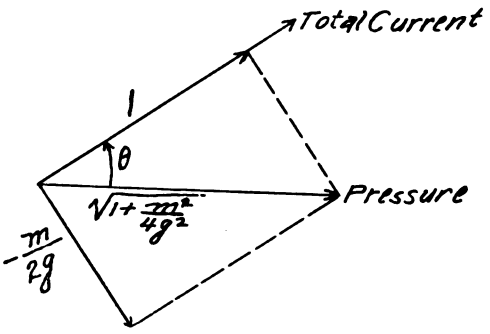

FIGURE 10

This occurs when $m^2=4K^2$ or at a frequency the wave length of which is: $\lambda=\frac{4}{m}$. If we take $m=0.07$ (a value used by Slepian and Hanna), then $\lambda=\frac{4\pi}{0.07}=180$ cm. This corresponds to a frequency of about 185 cycles. At lower frequencies the equations will not hold, because the power factor and power becomes imaginary. We can, however, increase the power output of these low frequencies by making m smaller. Such a procedure means that a horn must have a long neck. Also since σ_o the initial opening, should be small to increase the output as shown by equation (23), a small m means a long neck of smaller diameter. Such a procedure, if carried to extremes, would reduce the output to a large extent, particularly at high frequencies. As is well known, frequencies above 500 cycles are far more important for clearness of reproduction than are those frequencies below 500 cycles. In our effort to secure a greater low frequency output in this way, we should be cautious in order not to go too far.

We can make a comparison by means of our equations among the three types of radiators considered thus far. We saw that the

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conical horn did not cease radiating until zero frequency was reached. At the low frequencies, then, the conical horn should exceed the exponential one, as far as the simple theory we have used would cover the facts. In this respect, too, the straight pipe also should be superior to the exponential shape at these low frequencies, because it continues to radiate down to zero frequency.

If we refer to equation (23), the sound radiation at the initial end of the exponential horn is:

$$P_{x=0} = \frac{\rho A^2 c \sqrt{K^2 - \frac{m^2}{4}}}{2\sigma_o K} \tag{25}$$

Since the velocity and pressure along the horn both vary as $\epsilon^{-\frac{mx}{2}}$ and the area varies as ϵ^{-mx}, this equation represents the total radiation along the horn and is seen to be independent of x. We may make use of equation (25) then, to study what happens under various conditions.

At frequencies sufficiently high where K^2 is large enough to make $\frac{m^2}{4}$ a negligible factor in comparison, equation (25) becomes

$$P_{x=0} = \frac{\rho A^2 c}{2\sigma_o} \tag{26}$$

This result is identical with that for the straight pipe as given by equation (16). Under this condition we may note also that $\cos\theta$ becomes unity (see equation (24)) and the current and pressure are in phase as was shown to be the case for the straight pipe. Under the same conditions, we saw that the same conclusions were reached for the conical horn except that the radiation increased as the square of the frequency, whereas with the pipe and exponential horn the radiation is independent of frequency at these higher frequencies. As an illustration let us use the same numerical figures for m ($=0.07$) as used above for the exponential horn. Suppose, further, that $\sigma_o = 1$ sq. cm. (perhaps too small a value for practical purposes, but a value which tends to favor theoretically the exponential horn) and that $\omega = 0.01$ radian. Such a conical horn would have a cross sectional area of 400 sq. cm. at a distance of 200 cm. from the vertex. A comparison of equations (6) and (26) shows that the ratio between the conical and exponential horns at relatively high frequencies is given by:

$$\frac{P_c}{P_e} = \frac{K^2\sigma_o}{\omega} \tag{27}$$

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Substitution of the chosen values in this equation will show that at all frequencies above 500 cycles, approximately, the conical horn is superior to the exponential horn. It is allowable to use equation (25) in making this comparison because at 550 cycles m^2 is negligible compared with K^2. If, therefore, $m = 0.07$, $\tau_o = 1$, $\omega = 0.01$ and the same source of sound is used on both horns, the conical would be superior to the exponential not only above 550 cycles, but also below 185 cycles as previously calculated. Between these two frequency limits the exponential horn will be superior according to our theory and the amount of superiority can be calculated from the ratio:

$$\frac{P_c}{P_{E_{x=0}}} = \frac{K^3 \sigma_o}{\omega \sqrt{K^2 - \frac{m^2}{4}}} \tag{28}$$

which is obtained from equations (6) and (25). These results are not in disagreement with those of Slepian and Hanna—they merely show that by slightly altering conditions, the superiority or inferiority of one type of horn compared with another can be greatly changed and the whole complexion of the problem is altered. We desire to point out these differences so that a better understanding of the problem will result and thus lead to improved types of horns.

The results and comparison given above apply only to the case where the source of sound is placed at the vertex of the cone. Hanna and Slepian's results apply particularly to the case where the initial and final openings of the cone are equal to those of the exponential horn. In both these cases our theoretical deductions are in entire agreement. Since, however, the simple theory does not take all factors into consideration, and since the cones with finite openings, other than zero, have optimum angles for maximum output, it seems that exponential data will throw much additional light on the subject.

For this purpose, therefore, two straight conical and exponential horns were constructed of heavy gauge galvanized iron so that the walls would not vibrate. The initial and final areas and the horn lengths were the same, the latter being about 120 cm. An ordinary standard loud speaker unit was used for the middle and high frequency ranges. Over the middle register, the exponential horn was distinctly superior in accordance with predicted results. Over the high frequency region, the two horns performed very much alike, which is also in accordance with the theoretical results of Hanna and Slepian.

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Since superior low frequencies can be obtained from a unit with a low natural period, one was made with a natural period of about 300 cycles to test the horns at these frequencies. In the region of 80 to 200 cycles, the sound output from the conical horn was many times that from the exponential horn. The cut-off frequency of the exponential horn was calculated to be 135 cycles. With this same low frequency unit, the exponential horn became superior at about 200 cycles. These results, then, are in quite satisfactory qualitative agreement with the theoretical results which have been published by Hanna and Slepian and ourselves.

As stated in the beginning of this section, the above results include a review of existing literature, and some of our own results on horns of infinite lengths. We now proceed to consider some problems of horns of finite lengths.

4. Theory of Exponential Horns of Finite Length

The above discussion and the work which Dr. Slepian and Mr. Hanna have presented apply to horns of and tubes of infinite length. In these cases the sound energy indicated can be readily calculated provided certain simplifying assumptions are made. We have given these calculations for the straight tube, the straight cone, and for the exponential horn. The equations for the air-current, pressure, phase displacement, and energy radiated from the straight pipes and cones of infinite length are in accordance with the published work of Lord Rayleigh. To this material there has been added the case of the infinite exponential horn as a result of Webster's work and the extensions discussed above.

However, since we are called upon in practice to deal with horns and tubes of finite length, the extension of the theory for horns of unlimited lengths needs to be carried out to cover the cases of practical importance. As has been pointed out above, the foundation for this extension was laid by Professor A. G. Webster and published in general form with an application in condensed form to the special case of the straight pipe, the cone, the exponential horn, and the hyperbolic horn. These results were published in the "Proceedings for the National Academy of Sciences," pages 275-282, 1919. Professor G. W. Stewart[8] extended somewhat Webster's theory for the cone of limited length to include spherical waves and secured an equation which gave the ratio between the pressures at the vertex of the cone and those just outside the large opening of the horn in terms of its length and the pitch of the sound.

[8] Previous citation.

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We have extended Webster's theory for the exponential horn of finite length and have secured among others an equation for such a horn giving the same pressure ratio in terms of length, rate of opening up, and frequency as obtained by Professor Stewart for the conical horn. Since the ear is a pressure-responding device and not a power-responding device, the results in pressure changes are perhaps the more useful of the two possibilities. The following presentation of the theoretical results obtained and a comparison of the exponential and the conical horns will be of interest.

For this purpose we start with Webster's equations on page 282 of his article. These equations correspond to equation (11) (or the more general equation (13) with the time factor) in Appendix I. The time factor may be omitted for our purpose for the present. If we correct Webster's equation for typographical errors and use the quantities which we have already defined, then the pressure equation at the two ends and at any intermediate place may be written as follows:

$$\left.\begin{aligned} P_1 &= \epsilon^{-\frac{m x_1}{2}} [A \cos g x_1 + B \sin g x_1] \\ P_2 &= \epsilon^{-\frac{m x_2}{2}} [A \cos g x_2 + B \sin g x_2] \\ \text{and} \quad P &= \epsilon^{-\frac{m x}{2}} [A \cos g x + B \sin g x] \end{aligned}\right\} \tag{29}$$

These equations, when combined with the time factor, include propagation in both directions. As shown in Appendix I, $A = A_1 + B_1$ and $B = (A_1 - B_1)\, i$ where A_1 and B_1 represent the maximum pressures in the direct and reflected wave, respectively. The method of analysis in working with these exponential equations is the same as that used by Webster in his work on conical horns. Since our results on exponential horns have not been published and since they are of considerable interest, we believe it is of value to present a summary of them at this time. As shown in Appendix II, we start with the above equations and derive equations for pressures, displacements, impedances, the coefficients A and B, A_1 and B_1, etc. These equations are given here rather in summary form. They are worked out in detail in the Appendix II.

Professor Webster defined acoustic impedance by:

$$Z = \frac{P}{X} \tag{30}$$

rather than by P divided by $\frac{dx}{dt}$ according to the electrical

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analogy. X corresponds to the total displacement of fluid. For the general case Webster had for the acoustic impedance of the two ends of a horn:

$$Z_1 = \frac{Z_2 d - b}{-Z_2 c + a} \text{ and } Z_2 = \frac{Z_1 a + b}{Z_1 c + d} \tag{31}$$

We make use of equations (29) to determine the constants a, b, c, and d for the exponential horn and thus arrive at the expression for the acoustic impedances at the two ends of the horn. The proper adjustment of the impedances, according to Webster's viewpoint, at the two ends of the horn determine its efficiency in performing its functions of helping the radiation of sound from vibrating body, such as a diaphragm.

The mathematical work involved in determining the value of the constants a, b, c, and d is given in Appendix II. If the values given there are substituted in equations (31) we arrive at the following values for the acoustic impedances of the ends:

$$Z_1 = \frac{Z_2 \frac{\sigma_2}{\sigma_1}\left(\cos g L - \frac{m}{2g}\sin g L\right) - \frac{\beta K}{\sigma_1 g}\sin g L}{Z_2 \frac{K\sigma_2}{g\beta}\sin g L + \cos g L + \frac{m}{2g}\sin g L} \tag{32}$$

$$Z_2 = \frac{Z_1\left(\cos g L + \frac{m}{2g}\sin g L\right) + \frac{\beta K}{\sigma_1 g}\sin g L}{-Z_1 \frac{K\sigma_2}{g\beta}\sin g L + \frac{\sigma_2}{\sigma_1}\left(\cos g L - \frac{m}{2g}\sin g L\right)} \tag{33}$$

All the quantities have been defined in the previous work. These equations are similar to those for the impedance of conical horns given by Professor Webster. A discussion of these equations and a comparison of them with the equations for the conical horns is deferred until the additional equations are derived.

If X_1 and X_2 represent the total air displacement at the small and large end of the horn, respectively, then, as shown by Webster:

$$P_2 = a P_1 + b X_1 \text{ and } X_2 = c P_1 + d X_1 \tag{34}$$

If we substitute the values for a, b, c, and d determined in Appendix II in these equations and place $X_1 = 0$ and $X_2 = L$, then the pressure at the large end is given by*:

*In this and the following equations we may place

$$\varepsilon^{\frac{mL}{2}} = \sqrt{\frac{\sigma_2}{\sigma_1}} \text{ and } \varepsilon^{-\frac{mL}{2}} = \sqrt{\frac{\sigma_1}{\sigma_2}}$$

if we so desire.

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$$P_2=\epsilon^{-\frac{mL}{2}}\left[\cos g L+\frac{m}{2g}\sin g L\right]P_1+\frac{K\beta\epsilon^{-\frac{mL}{2}}}{g\sigma_1}X_1\sin g L \quad (35)$$

And the displacement at the large end is given by:

$$X_2=-\frac{P_1 K\sigma_2\epsilon^{-\frac{mL}{2}}}{g\beta}\sin g L+\frac{\sigma_2}{\sigma_1}\epsilon^{-\frac{mL}{2}}\left[\cos g L-\frac{m}{2g}\sin g L\right]X_1 \quad (36)$$

A discussion of these two equations is deferred also until additional relations are obtained for the exponential horn.

In Appendix II there has been derived the following equation which shows the relation between the pressure, P_1, at the small end of the exponential horn and the pressure, P_2, just outside the large end of the horn:

$$P_1=\frac{P_2\epsilon^{\frac{mL}{2}}}{\cos g L+\frac{m}{2g}\sin g L+\frac{\sigma_2 KZ_a}{g\beta}\sin g L} \quad (37)$$

In this equation L is the axial length of the exponential horn, σ_2 is the area of the large opening, Z_a is the acoustic impedance of the opening treated as a fictitious cylinder of definite length and an opening equal to the diameter of the large end. The equation for Z_a is derived in the Appendiz II, equation (19). $\beta=\rho C K^2$ where ρ is the air density, C is the velocity of sound and $K=2\pi$ divided by the wave length. g is defined by $\frac{1}{2}\sqrt{4K^2-m^2}$. m is defined by the exponential area, σ, thus, $\sigma=\sigma_1\epsilon^{mx}$. Our treatment of the opening is the same as developed by Rayleigh. Both Webster and Stewart used the same scheme in their theoretical work on horns.

To study the direct and reflected waves separately and in combination we make use of the equations for A and B, and A_1 and B_1 derived in Appendix II. The following equations are derived for these four constants:

$$A=-P_1 \text{ and } B=\frac{P_1\epsilon^{-\frac{mL}{2}}\cos g L-P_2}{\epsilon^{\frac{mL}{2}}\sin g L}$$

$$A_1=\frac{1}{2}\left(-P_1+\frac{P_1 i\epsilon^{-\frac{mL}{2}}\cos g L-P_2 i}{\epsilon^{-\frac{mL}{2}}\sin g L}\right) \quad (38)$$

and

$$B_1=\frac{1}{2}\left(-P_1-\frac{P_1 i\epsilon^{-\frac{mL}{2}}\cos g L-P_2 i}{\epsilon^{-\frac{mL}{2}}\sin g L}\right)$$

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In our study with the direct and reflected waves we are to work with A_1 and B_1 and not with A and B. The two former quantities represent the maximum pressures in the direct and reflected pressure waves of the type A

$$A_1 \varepsilon^{-\frac{mL}{2}} \cos(nt-gx) \text{ and } B_1 \varepsilon^{-\frac{mL}{2}} \cos(nt+gx)$$

respectively. These equations hold only for the case where $x=0$ and $x_2=L$, that is, the zero of the abscissas begin at $x=0$.

Since the pressure, P_1 equals $\rho a^2 s$ where s is the compression $\left(\text{which is equal to } \frac{dq}{dx}\right)$, we can make use of the pressure equations (29) as shown in Appendix II, to obtain the following expression for the total air displacement X, at any point in the horn:

$$X=-\frac{\sigma g \varepsilon^{-\frac{mx}{2}}}{\rho C^2 K^2}\left[A\left(\sin gx+\frac{m}{2g}\cos gx\right)-B\left(\sin gx-\frac{m}{2g}\cos gx\right)\right] \tag{39}$$

We are now in a position to make a comprehensive study of the exponential horn of finite length and compare the performance of this horn with that of the conical horn.

Discussion of Theory of Exponential Horn of Finite Length and Comparison to a Similar Conical Horn and Straight Pipe.

First, let us collect the equations we have derived for the exponential horn. We consider $X=0$ and $X_2=L$ for this summary.

Acoustic impedances:

Small end:

$$Z_1=\frac{Z_2\left[\frac{\sigma_2}{\sigma_1}\left(\cos gL-\frac{m}{2g}\sin gL\right)\right]-\frac{\beta k}{\sigma_1 g}\sin gL}{Z_2\frac{\sigma_2 K}{g\beta}\sin gL+\cos gL+\frac{m}{2g}\sin gL} \tag{32}$$

and for the large end:

$$Z_2=\frac{Z_1\left(\cos gL+\frac{m}{2g}\sin gL\right)+\frac{\beta k}{\sigma_1 g}\sin gL}{-Z_1\frac{K\sigma_2}{g\beta}\sin gL+\frac{\sigma_2}{\sigma_1}\left(\cos gL-\frac{m}{2g}\sin gL\right)} \tag{33}$$

Pressure and displacement at the large end in terms of corresponding quantities at small end:

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$$P_2 = \varepsilon^{-\frac{mL}{2}} \left[P_1 \left(\cos g L + \frac{m}{2g} \sin g L \right) + \frac{K\beta X_1}{g\sigma_1} \sin g L \right] \quad (35)$$

and

$$X_2 = \epsilon^{-\frac{mL}{2}} \left[\frac{\sigma_2 P_1 K}{\beta g} \sin g L + \frac{\sigma_2 X_1}{\sigma_1} \left(\cos g L - \frac{m}{2g} \sin g L \right) \right] \quad (36)$$

Pressure at small end in terms of pressure just outside large end:

$$P_1 = \frac{P_3 \varepsilon^{\frac{mL}{2}}}{\cos g L + \frac{m}{2g} \sin g L + \frac{K\sigma_2 Z_0}{g\beta} \sin g L} \quad (37)$$

For the pressure as a function of x we had:

$$P = \varepsilon^{-\frac{mx}{2}} (A \cos g x + B \sin g x) \quad (29)$$

where, as shown in Appendix II,

$$\left.\begin{aligned} A &= -P_1 \text{ and } B = -\frac{P_1 \varepsilon^{-\frac{mL}{2}} \cos g L - P_2}{\varepsilon^{-\frac{mL}{2}} \sin g L} \\ A_1 &= \frac{1}{2}\left(-P_1 + \frac{P_1 i \varepsilon^{-\frac{mL}{2}} \cos g L - {}_2 P i}{\varepsilon^{-\frac{mL}{2}} \sin g L} \right) \end{aligned}\right\} \quad (38)$$

for the direct waves and for the reflected waves:

$$B_1 = \frac{1}{2}\left(-P_1 - \frac{P_1 i \varepsilon^{-\frac{mL}{2}} \cos g L - P_2 i}{\varepsilon^{-\frac{mL}{2}} \sin g L} \right)$$

Finally the displacement as a function of x is shown by equation (5′) in Appendix II:

$$X = -\frac{\sigma g}{\rho C^2 K^2} \varepsilon^{-\frac{mx}{2}} \left[A \left(\sin g x + \frac{m}{2g} \cos g x \right) + B \left(\sin g x - \frac{m}{2g} \cos g L \right) \right] \quad (39)$$

The general equations (32) and (33) for the impedances may be considered from the practical importance of loud speakers. For this case the small end of the horn is closed. No energy, therefore, will be transmitted beyond the diaphragm closing the horn. A good method to employ in order to demonstrate that the receiver and the end of the horn is closed is to excite a long horn with a low frequency current thru the unit. The sound output will be quite marked if the frequency corresponds to one of the natural frequencies of the horn. If a small hole or crack is introduced at the connection between the horn and the receiver case,

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the low frequency sound output practically vanishes. The horn is now acting like one with both ends opened. It is astonishing to find that so small an opening will cause so marked an effect. The practical importance of exercising care in this direction is apparent.

We may follow Webster and Stewart, therefore, by placing the impedance, Z_1, at the small end of the horn equal to infinity. The impedance at the large end, then, can be written from equation (33) as follows:

$$Z_2 = -\frac{\cos g L + \frac{m}{2g}\sin g L}{\frac{K\sigma_2}{\beta g}\sin g L} \tag{40}$$

Put $g = \frac{1}{2}\sqrt{4K^2 - m^2}$ and $\beta = \rho C^2 K$. We get, then,

$$Z_2 = \frac{\rho C^2}{2\sigma_2}\left[\sqrt{4K^2 - m^2}\cot\frac{L}{2}\sqrt{4K^2 - m^2} + m\right] \tag{41}$$

We observe that the impedance of the large end varies inversely as its area. Since the acoustic impedance of the large end should be as small as possible in order to have a minimum reflection loss, it is necessary to have a large final opening. The above equation is a general one to cover the whole frequency range. We may simplify it if we desire to study special cases. For example, if we consider m^2 small in comparison with $4K^2$ as it will be for frequencies above about 1,000 cycles for practical values of m, then we may write equation (41) as follows:

$$\underset{\text{high freq.}}{Z_2} = -\frac{\rho C^2}{2\sigma_2}\left[2K\cot KL + m\right] \tag{42}$$

When $KL = \frac{n_2\pi}{2}$ where $n_2 = 2, 4, 6$, etc., we have $\cot KL = \infty$;

and when $KL = \frac{n_1\pi}{2}$ where $n_1 = 1, 3, 5, 7$, etc., we have $\cot KL = 0$.

The former case corresponds to high frequency anti-resonance where Z_2 is very large. The latter corresponds to:

$$Z_2 = \frac{\rho C^2 m}{2\sigma_2} \tag{43}$$

which corresponds to high frequency resonance where Z_2 has its smallest value. The smallest value, then, is limited by the value of m and that of the final area. There is, then, even at these high frequencies a wide variation in the acoustic impedance of the exponential horn.

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The special case where $\frac{1}{2}\sqrt{4K^2-m^2}$ approaches zero, which occurs at some definite low frequency, is of interest. $cot\frac{L}{2}\sqrt{4K^2-m^2}$ can be replaced by $\frac{2}{L\sqrt{4K^2-m^2}}$. Substitute this in (41) and we get:

$$Z_2=\frac{\rho C^2}{2\sigma_2}\left[\frac{2}{L}+m\right] \tag{44}$$

Consequently, at these low frequencies near the critical frequency the acoustical impedance is a constant finite value and is determined by the final opening, length, and exponential coefficient, m, of the horn.

A comparison between the impedances for a straight pipe, the exponential, and the conical horns is of interest. Webster[9] showed that the impedance for a straight pipe is

$$Z_2=\frac{\beta}{\sigma}\left(\frac{Z_1 cos\,K\,l+\frac{\beta}{\sigma}sin\,K\,l}{-Z_1 sin\,K\,l+\frac{\beta}{\sigma}cos\,K\,l}\right) \tag{45}$$

If Z_1 is infinity, as in the case of the pipe closed at the end, then, when β is replaced by $\rho C^2 K$

$$Z_2=-\frac{\rho C^2 K}{\sigma}cot\,Kl \tag{46}$$

This is the same as equation (42), provided m is zero and $\sigma=\sigma_2$. When m is zero, the exponential horn reduces to a straight pipe and the reduction of equation (42) for the exponential horn to the form for a pipe as given by Webster, speaks for the correctness of our equations. Z_2 for both the pipe and the exponential with a small m varies between wide limits corresponding to resonance and anti-resonance. In both types of apparatus, at the anti-resonance frequencies, the value of Z_2 becomes larger with increasing frequencies because of the factor K. This would tend to decrease the sound radiation more and more the higher the frequency.

A comparison of the acoustic impedance of the large end of the conical horn can be made with that for the exponential horn also. The acoustic impedance for the large end of the conical horn when the opposite end is closed by a receiver diaphragm is:[10]

$$Z_2=-\frac{\rho C^2 K}{\sigma_2}\left[\frac{sin\,K\,(L+\varepsilon_1)\,sin\,K\,\varepsilon_2}{sin\,K\,(L+\varepsilon_1-\varepsilon_2)}\right] \tag{47}$$

[9] Previous citation 4.
[10] Previous citation 4.

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This equation is derived from the first equation for Z_2 on page 280 of Webster's article by placing $Z_1 = \infty$. In this equation ϵ_1 and ϵ_2 are defined by $tan\, K\epsilon_1 = k\, x_1$ and $tan\, K\epsilon_2 = K x_2$. If $x_1 = 0$ then $x_2 = L$, the length of the horn. Also, if $x_1 = 0$, then $tan\, K\epsilon_1 = 0$, and therefore $\epsilon_1 = 0$. When $x_2 = L$, $tanK\epsilon_2 = KL$ or $K\epsilon_2 = tan^{-1}KL$. Using these relations, (47) reduces to:

$$Z_2 = -\frac{\rho C^2 K}{\sigma_2}\left[\frac{sinKLsin(tan^{-1}KL)}{sin(KL+tan^{-1}KL)}\right] \tag{48}$$

The corresponding equation for the exponential horn is given by equation (41). The final opening of the two horns enter the equations the same way. The impedance of the conical horn and the straight pipe increases directly with the frequency; that of the exponential varies in practically the same way. Hence, at high frequencies the reflection losses will be large, due to too high an acoustic impedance for the three types of devices considered here,—each performing in essentially the same manner. Because of this, the horns are more efficient the lower the frequency.

We shall next consider equation (37). It may be written:

$$P_3 = P_1 \varepsilon^{-\frac{mL}{2}}\left[cos\frac{1}{2}\sqrt{4K^2-m^2}\,L + \frac{m}{\sqrt{4K^2-m^2}}sin\frac{1}{2}\sqrt{4K^2-m^2}\,L + \frac{2\sigma_2 KZ_o}{\beta\sqrt{4K^2-m^2}}sin\frac{1}{2}\sqrt{4K^2-m^2}\,L\right] \tag{49}$$

The impedance, Z_o, of the large end treated as a tube of finite length radiating into the outside space as given by equation (19) in Appendix II is:

$$Z_o = \rho C^2 K^2\left(\frac{Ki}{2\pi} - \frac{1}{C_o}\right) \tag{50}$$

In this equation C_o is the acoustic conductance of the opening as treated and defined in connection with the development of this equation in Appendix II. i is the square root of (minus one), and merely means that the vector sum of the two terms of equation (50) should be taken. The absolute value of Z_o would need to be used in equation (49). To use this equation we would need to know the value of C_o, the acoustic conductance of the open end treated in the manner stated. If R is the radius of the open end, then Rayleigh on page 181, volume 2, "Theory of Sound," gives

$$C_o = \frac{\pi R^2}{\frac{1}{4}\pi R}.$$

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The special case where $\frac{1}{2}\sqrt{4K^2-m^2}$ approaches zero, which occurs at some definite low frequency, is of interest. $cot\frac{L}{2}\sqrt{4K^2-m^2}$ can be replaced by $\frac{2}{L\sqrt{4K^2-m^2}}$. Substitute this in (41) and we get:

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Consequently, at these low frequencies near the critical frequency the acoustical impedance is a constant finite value and is determined by the final opening, length, and exponential coefficient, m, of the horn.

A comparison between the impedances for a straight pipe, the exponential, and the conical horns is of interest. Webster[9] showed that the impedance for a straight pipe is

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If Z_1 is infinity, as in the case of the pipe closed at the end, then, when β is replaced by $\rho C^2 K$

$$Z_2=-\frac{\rho C^2 K}{\sigma}cot\,Kl \tag{46}$$

This is the same as equation (42), provided m is zero and $\sigma=\sigma_2$. When m is zero, the exponential horn reduces to a straight pipe and the reduction of equation (42) for the exponential horn to the form for a pipe as given by Webster, speaks for the correctness of our equations. Z_2 for both the pipe and the exponential with a small m varies between wide limits corresponding to resonance and anti-resonance. In both types of apparatus, at the anti-resonance frequencies, the value of Z_2 becomes larger with increasing frequencies because of the factor K. This would tend to decrease the sound radiation more and more the higher the frequency.

A comparison of the acoustic impedance of the large end of the conical horn can be made with that for the exponential horn also. The acoustic impedance for the large end of the conical horn when the opposite end is closed by a receiver diaphragm is:[10]

$$Z_2=-\frac{\rho C^2 K}{\sigma_2}\left[\frac{sin\,K\,(L+\varepsilon_1)\,sin\,K\,\varepsilon_2}{sin\,K\,(L+\varepsilon_1-\varepsilon_2)}\right] \tag{47}$$

[9] Previous citation 4.
[10] Previous citation 4.

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This equation is derived from the first equation for Z_2 on page 280 of Webster's article by placing $Z_1 = \infty$. In this equation ϵ_1 and ϵ_2 are defined by $tan\, K\epsilon_1 = k\, x_1$ and $tan\, K\epsilon_2 = K x_2$. If $x_1 = 0$ then $x_2 = L$, the length of the horn. Also, if $x_1 = 0$, then $tan\, K\epsilon_1 = 0$, and therefore $\epsilon_1 = 0$. When $x_2 = L$, $tan K\epsilon_2 = KL$ or $K\epsilon_2 = tan^{-1} KL$. Using these relations, (47) reduces to:

$$Z_2 = -\frac{\rho C^2 K}{\sigma_2}\left[\frac{sin KL\, sin(tan^{-1} KL)}{sin(KL + tan^{-1} KL)}\right] \tag{48}$$

The corresponding equation for the exponential horn is given by equation (41). The final opening of the two horns enter the equations the same way. The impedance of the conical horn and the straight pipe increases directly with the frequency; that of the exponential varies in practically the same way. Hence, at high frequencies the reflection losses will be large, due to too high an acoustic impedance for the three types of devices considered here,—each performing in essentially the same manner. Because of this, the horns are more efficient the lower the frequency.

We shall next consider equation (37). It may be written:

$$P_3 = P_1 \varepsilon^{-\frac{mL}{2}}\left[cos\frac{1}{2}\sqrt{4K^2 - m^2}\,L + \frac{m}{\sqrt{4K^2 - m^2}} sin\frac{1}{2}\sqrt{4K^2 - m^2}\,L + \frac{2\sigma_2 K Z_o}{\beta\sqrt{4K^2 - m^2}} sin\frac{1}{2}\sqrt{4K^2 - m^2}\,L\right] \tag{49}$$

The impedance, Z_o, of the large end treated as a tube of finite length radiating into the outside space as given by equation (19) in Appendix II is:

$$Z_o = \rho C^2 K^2\left(\frac{K i}{2\pi} - \frac{1}{C_o}\right) \tag{50}$$

In this equation C_o is the acoustic conductance of the opening as treated and defined in connection with the development of this equation in Appendix II. i is the square root of (minus one), and merely means that the vector sum of the two terms of equation (50) should be taken. The absolute value of Z_o would need to be used in equation (49). To use this equation we would need to know the value of C_o, the acoustic conductance of the open end treated in the manner stated. If R is the radius of the open end, then Rayleigh on page 181, volume 2, "Theory of Sound," gives

$$C_o = \frac{\pi R^2}{\frac{1}{4}\pi R}.$$

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Experimental data (Rayleigh, page 202) indicated that

$$C_o = \frac{\pi R}{0.6}$$

which we can use in equation (50) to determine Z_o.

At the higher frequencies, m^2 can be neglected compared with $4K^2$. If $m^2 = 0.005$, then when $4K^2 = 0.05$ or $K = 0.11$, corresponding to about 600 cycles. Equation (49) can be written:

$$P_3 = \underset{\text{high freq.}}{P_1 \epsilon^{-\frac{mL}{2}}} \left[\cos KL + \left(\frac{m}{2K} + K\sigma_2 \left(\frac{Ki}{2\pi} - \frac{1}{C_o} \right) \right) \sin KL \right] \qquad (51)$$

To get this equation from (49) we made the following substitution

$$\frac{2K\sigma_2 Z_o}{\beta\sqrt{4K^2 - m^2}} = K\sigma_2 \left(\frac{Ki}{2\pi} - \frac{1}{C_o} \right)$$

which is legitimate for the high frequencies which we are considering. In a typical example of practical value, $\frac{1}{C_o}$ will be negligible compared with $\frac{K}{2\pi}$, σ_2 will be perhaps 2,000 sq. cm., in which case $\frac{m}{2K}$ will be negligible compared with $\frac{\sigma_2 K^2}{\pi}$.

Introducing these restrictions in (51) we get:

$$P_3 = P_1 \epsilon^{-\frac{mL}{2}} \left[\cos KL + \frac{\sigma_2 K^2}{\pi} \sin KL \right] \qquad (52)$$

The coefficient of the second term is large compared with that of the cosine term and the pressure just outside the horn shows large increase over that at the small end due to the phenomenon of resonance. The minimum value of P_3 will be determined by $\cos KL$, and will occur at frequencies corresponding to 2, 4, 6, etc., times the fundamental and the maximum value of P_3 will occur at frequencies 1, 3, 5, 7, etc., times the fundamental. In this respect the pressure outside the large end of the exponential horn follows the changes in acoustic impedance of the large end but in the reverse direction. That is, increased pressure occurs with decrease acoustic impedance and vice versa.

At the lower frequencies we must use the more general relation given in equation (49). This equation shows the same resonant and anti-resonant frequencies as did the restricted equations (51) and (52). In the latter case, however, there was a harmonic relation of the resonant frequencies. In the former case, however, the maxima and minima of the sine and cosine

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terms occur when $\frac{1}{2}\sqrt{4K^2-m^2}\,L=\frac{n\pi}{2}$, where n is 0, 1, 2, 3, 4, etc. This relation gives $\lambda=\frac{4\pi L}{\sqrt{n^2\pi^2+m^2L^2}}$ for the maxima and minima of the sine and cosine terms of equation (49).

The special case of $n=0$, that is, $\sqrt{4K^2-m^2}=0$ is of interest and has been discussed in connection with the impedance of the open end. We also encountered this case for the exponential horn of infinite length where the phase angle between the current and pressure became $\frac{\pi}{2}$ and the power radiation became zero. In the case of the pressures under discussion here, when $\sqrt{4K^2-m^2}$ approaches and finally becomes zero, all the terms in (49) remain finite and the pressures are therefore finite. When, however, $\sqrt{4K^2-m^2}$ becomes imaginary, as it does at low frequencies, then equation (49) becomes

$$P_3=P_1\varepsilon^{-\frac{mL}{2}}\left[\cosh\frac{1}{2}\sqrt{m^2-4K^2}L+\frac{m}{\sqrt{m^2-4K^2}}\sinh\frac{1}{2}\sqrt{m^2-4K^2}L+\frac{2\sigma_2 Z_o K}{\beta\sqrt{m^2-4K^2}}\sinh\frac{1}{2}\sqrt{m^2-4K^2}L\right] \quad (53)$$

To arrive at this relation we substituted $\sin ix=i\sinh x$, $\cos ix=\cosh x$ and $\sqrt{4K^2-m^2}=i\sqrt{m^2-4K^2}$. We see, therefore, that the pressure, P_3 remains finite at these low frequencies, but varies continuously. That is, there are no resonant effects for frequencies below the critical frequencies. At the very low frequencies $4K^2$ can be neglected compared with m^2. In this case for a moderately long exponential horn with an m equal to about 0.07 the hyperbolic sines and cosines are practically equal. The coefficient of the second term reduces to unity, in which case the first two terms combine, but their coefficient is negligible compared with that of the third term. This leaves:

$$P_3=P_1\varepsilon^{-\frac{m}{2}2}\frac{2\sigma_2K^2}{m}\left(\frac{Ki}{2\pi}-\frac{1}{C_o}\right)\sinh^{\frac{mL}{2}} \quad (54)$$

where, for

$$\frac{KZ_o}{\beta\sqrt{m^2-4K^2}}$$

we are permitted for this case to substitute

$$\frac{K^2}{m}\left(\frac{Ki}{2\pi}-\frac{1}{C_o}\right)$$

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This same equation (54) also is sufficiently accurate at the critical frequency because the first two terms are small in comparison with the third one. Consequently the variation in P_3 with frequency at these low values of frequency is determined by the change introduced because of K in equation (54). The lowest limit of P_3 is obtained when K is placed equal to zero, when the pressure P_3 becomes equal to zero.

As stated earlier in the paper, we are interested mainly in pressure changes because the ear itself responds to pressure changes in the air and also because we are able to make experimental tests on such pressure changes. For these reasons in the discussion of our theoretical work we have given much space to studying these effects. Furthermore, the amount of sound energy radiated into the surrounding space is directly associated with the acoustic impedances of the large end of the horn. It is for this reason that much consideration has been given to a study of acoustic impedances.

It is of interest to compare Professor Stewart's pressure equation for the conical horn with our equation for the exponential horn. He had (equation 2, page 322, "Physical Review," 1920):

$$P_1 = \frac{P_3}{\frac{\sin Kr}{Kr} + \frac{\sigma_2 \sin K(r-\varepsilon)}{r \sin K\varepsilon}\left(\frac{Ki}{2\pi} - \frac{1}{C_o}\right)} \tag{55}$$

where $\tan \epsilon = K r$, r being the length of the horn. The other quantities involved in this equation are the same as used by us to derive equation (37). Reference may be made to equation (19) in Appendix II to see that we use the same expression for Z_o as used by Professor Stewart. Equation (55) shows the same type of resonant phenomenon as does our equation (37). In the conical horn equation, however, there is no "critical" frequency as found for the exponential case; neither is there an exponential factor as obtained in the exponential pressure equation. At very low frequencies (much below the fundamental of the horn) equation (55) can be replaced by:

$$P_3 = P_1\left[1 + \frac{\sigma_2(r-\varepsilon)}{r\varepsilon}\left(\frac{Ki}{2\pi} - \frac{1}{C_o}\right)\right] \tag{56}$$

which corresponds to equation (54).

But, for these small values of Kr, $K\varepsilon = \tan^{-1} Kr$ can be replaced by $K\varepsilon = Kr$ or $\varepsilon = r$. Whence at all the very low frequency region where this condition is satisfied, equation (56) states that $P_3 = P_1$ and is independent of the horn. This is not the case for the exponential horn as shown above where $P_3 = 0$ at zero frequency.

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At the high frequencies where Kr is large (say for all frequencies above 1,000 cycles), where r is about 200 cm., equation (55) can be modified. The term $\frac{\sin Kr}{Kr}$ can be neglected. Since Kr is large, $\tan K\epsilon$ is large, and, therefore, $K\epsilon$ can be placed equal to $\frac{\pi}{2}$. In this case $\sin K\epsilon$ is essentially unity. Also, since $K\epsilon = \frac{\pi}{2}$ it means that ϵ is about 7 or 8 at 1,000 cycles and becomes less with increasing frequencies. Consequently, if r is 200 cm. $(r-\epsilon)$ can be placed equal to r. With these simplifications equation (55) can be written

$$\frac{P_3}{P_1} = \frac{\sigma_2 \sin Kr}{r}\left(\frac{Ki}{2\pi} - \frac{1}{C_o}\right) \tag{57}$$

The higher the frequency, the more accurate this equation becomes, finally taking on the form,

$$\frac{P_3}{P_1} = \frac{\sigma_2 K \sin Kr}{2\pi r} \tag{58}$$

because of the relative smallness of $\frac{1}{C_o}$ at these high frequencies. Consequently, the resonant frequencies are harmonic and continue up to these higher frequencies. The performance, then, of the exponential and conical horns are quite similar at these high frequencies. A comparison of equation (52) with (58) shows both the similarities and differences at these high frequencies between the two types of horns.

A word as to the reflection effects and to the energy in the direct and reflected waves will also be of interest. A detail report of these effects is reserved for a future publication. For the present, however, we may call attention to the equations for the coefficient of the direct and reflected waves given in (38). From these equations the ratio of the pressure coefficients for the reflected and direct waves is:

$$\frac{B_1}{A_1} = \frac{P_1 \varepsilon^{-\frac{mL}{2}} [\sin gL + i\cos gL] - P_2 i}{P_1 \varepsilon^{-\frac{mL}{2}} [\sin gL - i\cos gL] + P_2 i} \tag{59}$$

But the absolute value of

$$\sin gL + i\cos gL = \sqrt{\sin^2 gL + \cos^2 gL} = 1$$

and the absolute value of

$$\sin gL - i\cos gL = \sqrt{\sin^2 gL + \cos^2 gL} = 1$$

so that:

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$$\frac{B_1}{A_1} = \frac{P_1 \varepsilon^{-\frac{mL}{2}} - P_2 i}{P_1 \varepsilon^{-\frac{mL}{2}} + P_2 i} \tag{60}$$

If equation (35) is substituted for P_2 in terms of P_1 it will be possible to make a comparative study of the ratio of reflected to direct pressures. Other equations may be developed, but these are sufficient to indicate the direction the investigation takes.

A study of energy radiations at the large end can be made by means of equations (35 and (36). To do this, it is first necessary to secure the total "current" from the total displacement X_2. This is done merely by using the $\frac{dx_2}{dt}$ instead of X_2. It is necessary first to put in the time factor in both equations (35) and (36) corresponding to equation (13) in Appendix I. The results of this study, however, will be presented with experimental data in a future publication to which we have just referred.

APPENDIX I

Webster's Derivation of Equations for Exponential Horns

The general equation for wave motion in three directions is given by the familiar equation:

$$\frac{d^2\phi}{dt^2} = C^2 \nabla^2 \phi = C^2 \left[\frac{d^2\phi}{dx^2} + \frac{d^2\phi}{dy^2} + \frac{d^2\phi}{dz^2} \right] \tag{1}$$

ϕ is the velocity potential for sound waves. For plane waves we use

$$\frac{d^2\phi}{dt^2} = C^2 \frac{d^2\phi}{dx^2} \tag{2}$$

To derive his general equations for horns of various shapes, Professor Webster introduced a variable cross section, σ, varying as a function of x. Introducing σ in (2) we get:

$$\frac{d^2\phi}{dt^2} = C^2 \left[\frac{1}{\sigma} \frac{d}{dx} \left(\sigma \frac{d\phi}{dx} \right) \right] \tag{3}$$

or

$$\frac{d^2\phi}{dt^2} = C^2 \left[\frac{d^2\phi}{dx^2} + \frac{1}{dx} \frac{d\sigma}{\sigma} \frac{d\phi}{dx} \right] \tag{4}$$

whence

$$\frac{d^2\phi}{dt^2} = C^2 \left[\frac{d^2\phi}{dx^2} + \frac{d(\log \sigma)}{dx} \frac{d\phi}{dx} \right] \tag{5}$$

For the exponential horn:

$$\sigma = \sigma_1 \varepsilon^{mx} \tag{6}$$

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so that

$$\frac{d}{dx}(\log \sigma) = m \tag{7}$$

Introduce this in (5) and we get:

$$\frac{d^2\phi}{dt^2} = C^2\left[\frac{d^2\phi}{dx^2} + m\frac{d\phi}{dx}\right] \tag{8}$$

For potentials of the harmonic type, such as $\phi \propto \varepsilon^{int}$ we have $\frac{d^2\phi}{dt^2} = -n^2\phi$. With this change equation (8) becomes

$$\frac{d^2\phi}{dx^2} + m\frac{d\phi}{dx} + K^2\phi = 0 \tag{9}$$

where $K^2 = \frac{n^2}{C_2} = \left(\frac{2\pi}{\lambda}\right)^2$ where λ equals the wave length.

To solve this equation we assume the ϕ is proportional to ε^{ax} and substitute in (9). We get two roots – determined by $a^2 + m\,a + K^2 = 0$—for a; namely

$$a_1 = -\frac{m}{2} + \frac{i}{2}\sqrt{4K^2 - m^2} \text{ and } a_2 = -\frac{m}{2} - \frac{i}{2}\sqrt{4K^2 - m^2}.$$

For propagation in both directions, therefore, the solution of (9) is:

$$\phi = A_1\varepsilon^{a_1 x} + B_1\varepsilon^{a_2 x} \tag{10}$$

or, in trigonometric form:

$$\phi = \varepsilon^{-\frac{mx}{2}}[A\cos g\,x + B\sin g\,x] \tag{11}$$

where $A = A_1 + B_1$, $B = (A_1 - B_1)i$ and $g = \frac{1}{2}\sqrt{4K^2 - m^2}$.

We should observe that A_1 and B_1 take care of the direct and reflected waves, respectively, but A and B do not. However, we may introduce the time factor in (10) and thus obtain the general solution of equation (8). Thus

$$\phi = \varepsilon^{-\frac{mx}{2}}[A_1\varepsilon^{igx} + B_1\varepsilon^{-igx}]\varepsilon^{int} \tag{12}$$

Remembering that:

$$\cos a\,x + i\sin a\,x = \varepsilon^{iax} \text{ and } \cos a\,x - i\sin a\,x = \varepsilon^{-iax},$$

we get for (12) where only the real portion is retained:

$$\phi = \varepsilon^{-\frac{mx}{2}}[A_1\cos(nt - gx) + B_1\cos(nt + gx)] \tag{13}$$

The equation (11) was published by Webster in his article referred to. It is only a step to equation (13). The first factor takes care of the wave motion in the outward direction, while

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the second term represents the reflected wave. A comparison of A_1 and B_1, therefore, enables us to study the reflected and direct wave. When, therefore, we considered the propagation in one direction we studied a special case of Professor Webster's more general equations. This, of course, is what we (and Slepian and Hanna, independently) have done in section 2 of the paper.

APPENDIX II

Analysis of the Exponential Horn Based on Webster's General Method

The equations given in Appendix I are entirely due to Webster. The equations giuen in this section, altho based on Webster's method of analysis, are new and of considerable importance in the study of horns. For this reason we believe it is worth while to present here a summary of this work.

If P, P_1, and P_2 are the sound pressures at any point along the horn, at the initial end and at the final or large end, then according to equation (11) of Appendix I, we may write:

$$P = A\,u + B\,v,\ P_1 = A\,u_1 + B\,v_1,\ \text{and}\ P_2 = A\,u_2 + b\,v_2 \qquad (1)$$

where $u\ (g\,x)$ and $v\ (g\,x)$ are independent solutions of equation (9) in Appendix I. The pressure, P, is equal to $-\rho\,C^2\frac{d\,q}{d\,x}$ where q is the displacement of air, ρ the density and C the velocity of sound. We can make use of this relation to determine equations for q. A comparison of (1) with equation (11) shows that

$$u = \varepsilon^{-\frac{m\,x}{2}} \cos g\,x \text{ and } v = \varepsilon^{-\frac{m\,x}{2}} \sin g\,x \qquad (2)$$

If we write, therefore,

$$p = -\rho\,C^2\frac{d\,q}{d\,x} = A\,\varepsilon^{-\frac{m\,x}{2}} \cos g\,x + B\,\varepsilon^{-\frac{m\,x}{2}} \sin g\,x \qquad (3)$$

we obtain

$$-\rho\,C^2\,q = A\int \varepsilon^{-\frac{m\,x}{2}} \cos g\,x\,d\,x + B\int \varepsilon^{-\frac{m\,x}{2}} \sin g\,x\,d\,x \qquad (4)$$

whence

$$\begin{aligned}\rho\,C^2\,q = &-\frac{A\,g}{K^2}\varepsilon^{-\frac{m\,x}{2}}\left[\sin g\,x + \frac{m}{2\,g}\cos g\,x\right] \\ &-\frac{B\,g}{K^2}\varepsilon^{-\frac{m\,x}{2}}\left[\sin g\,x - \frac{m}{2\,g}\cos g\,x\right]\end{aligned} \qquad (5)$$

Parenthetically, we may include the total displacement

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$X\,\sigma = q\,g$) by means of the following equation obtained by multiplying the two sides of (5) by σ and dividing by $\rho\,C^2$. Thus,

$$X = -\frac{g\,\sigma}{\rho\,C^2\,K^2}\,\varepsilon^{-\frac{m\,x}{2}}\left[A\left(\sin g\,x + \frac{m}{2\,g}\cos g\,x\right) + B\left(\sin g\,x - \frac{m}{2\,g}\cos g\,x\right)\right] \tag{5'}$$

On the other hand if we differentiate (2) with respect to ($g\,x$), (not x) we find that equation (5) can be written:

$$\left.\begin{aligned} \beta\,q &= A\,u^1 + B\,v^1 \\ \text{and therefore } \beta\,q_1 &= A\,u_1{}^1 + B\,v_1{}^1 \text{ and } \beta\,q_2 = A\,u_2{}^1 + B\,v_2{}^1 \end{aligned}\right\} \tag{6}$$

where $\beta = \rho\,C^2\,K$. Equations given in (1), (2), and (6) are used in developing the results which are to follow.

For convenience of manipulation the following determinants are defined and used:

$$\left.\begin{aligned} D_1 &= \begin{vmatrix} u_1 & v_1 \\ u_1{}^1 & v_1{}^1 \end{vmatrix}, \quad D_2 = \begin{vmatrix} u_2 & v_2 \\ u_2{}^1 & v_2{}^1 \end{vmatrix}, \quad D_3 = \begin{vmatrix} u_1 & v_1 \\ u_2{}^1 & v_2{}^1 \end{vmatrix} \\ D_4 &= \begin{vmatrix} u_2 & v_2 \\ u_1{}^1 & v_1{}^1 \end{vmatrix}, \quad D_5 = \begin{vmatrix} u_1 & v_1 \\ u_2 & v_2 \end{vmatrix}, \quad D_6 = \begin{vmatrix} u_1{}^1 & v_1{}^1 \\ u_2{}^1 & v_2{}^1 \end{vmatrix} \end{aligned}\right\} \tag{7}$$

With these definitions, the following four constants are defined:

$$a = \frac{D_4}{D_1}, \quad b = \frac{\beta\,D_5}{\sigma_1\,D_1}, \quad c = -\frac{\sigma_2\,D_6}{\beta\,D_1}, \quad d = \frac{\sigma_2\,D_3}{\sigma_1\,D_1} \tag{8}$$

where σ_1 represents the small opening and σ_2 the large one of the exponential horn defined by $\sigma = \sigma_1\,\varepsilon^{m\,x}$. Using the equations given by (2) we find from (7) that:

$$\left.\begin{aligned} D_1 &= \frac{g}{K}\,\varepsilon^{-m\,x_1}, \quad D_2 = \frac{g}{K}\,\varepsilon^{-m\,x_2} \\ D_3 &= \frac{g}{K}\,\varepsilon^{-\frac{m}{2}(x_2+x_1)}\left[\cos g(x_2-x_1) - \frac{m}{2\,g}\sin g(x_2-x_1)\right] \\ D_4 &= \frac{g}{K}\,\varepsilon^{-\frac{m}{2}(x_2+x_1)}\left[\cos g(x_2-x_1) + \frac{m}{2\,g}\sin g(x_2-x_1)\right] \\ D_5 &= \varepsilon^{-\frac{m}{2}(x_2+x_1)}\sin g(x_2-x_1), \quad D_6 = \varepsilon^{-\frac{m}{2}(x_2+x_1)}\sin g(x_2-x_1) \end{aligned}\right\} \tag{9}$$

Substitute these in (8) and we find:

$$\left.\begin{aligned} a &= \varepsilon^{-\frac{m}{2}(x_2-x_1)}\left[\cos g(x_2-x_1) + \frac{m}{2\,g}\sin g(x_2-x_1)\right] \\ b &= \frac{K\beta}{g\sigma_1}\,\varepsilon^{-\frac{m}{2}(x_2-x_1)}\sin g(x_2-x_1) \\ c &= -\frac{K\sigma_2}{g\beta}\,\varepsilon^{-\frac{m}{2}(x_2-x_1)}\sin g(x_2-x_1) \\ d &= \frac{\sigma_2}{\sigma_1}\,\varepsilon^{-\frac{m}{2}(x_2-x_1)}\left[\cos g(x_2-x_1) - \frac{m}{2\,g}\sin g(x_2-x_1)\right] \end{aligned}\right\} \tag{10}$$

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In the general case Professor Webster showed that the acoustic impedances of the two ends are given by:

$$\left.\begin{aligned} Z_1 &= \frac{Z_2 d - b}{-Z_2 c + a} \text{ for the small end} \\ \text{and} \quad Z_2 &= \frac{Z_1 a + b}{Z_1 c + d} \text{ for the large end} \end{aligned}\right\} \tag{11}$$

He derived the expressions for Z_1 and Z_2 for the conical horns. If we substitute our values of a, b, c, d, in equation (11) we arrive at the corresponding equations for the exponential horns. If we place $x_2 - x_1 = L$, the length of the horn, then we get for the impedances:

$$\left.\begin{aligned} Z_1 &= \frac{Z_2 \frac{\sigma}{\sigma_1}\left(\cos g L - \frac{m}{2g}\sin g L\right) - \frac{\beta K}{\sigma_1 g}\sin g L}{Z_2 \frac{K\sigma_2}{g\beta}\sin g L + \cos g L + \frac{m}{2g}\sin g L} \\ Z_2 &= \frac{Z_1\left(\cos g L + \frac{m}{2g}\sin g L\right) + \frac{\beta K}{\sigma_1 g}\sin g L}{-Z_1 \frac{K\sigma_2}{g\beta}\sin g L + \frac{\sigma_2}{\sigma_1}\left(\cos g L - \frac{m}{2g}\sin g L\right)} \end{aligned}\right\} \tag{12}$$

We now extend the analysis to secure a relation between the pressure, P_1, at the small end of the horn and that, P_3, just outside the large end. This analysis is similar to one used by Professor Stewart for the conical horn and published by him in the "Physical Review" for 1920, pages 313-326.

As the air passes from the end, X_2, to the surrounding space, it does so by effectively passing thru a short fictitious tube the length of which is ΔL, known as the end correction, and the diameter of which is that of the large end of the horn. At the initial end of this tube the pressure is P_2, and that at the other end is P_3. There is thus a change in pressure, $P_2 - P_3$, thru the length of this fictitious tube. If we call Z_o the acoustic impedance of this "tube," then by definition

$$X_2 Z_o = P_2 - P_3 = X_2 Z_2 - P_3 \tag{13}$$

where X_2 is the total displacement ($\sigma_2 q_2$) in the "tube," being the same as that at the end of the horn. From (13) we get

$$X_2 = \frac{P_3}{Z_2 - Z_o} \tag{14}$$

or

$$P_2 = \frac{P_3 Z_2}{Z_2 - Z_o} \tag{15}$$

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which is a general relation for all horns. In Webster's paper he showed that for the general case:

$$P_2 = a\,P_1 + b\,X_1 \tag{16}$$

which we apply to the exponential horn by using the values determined above for the exponental case. From (15) and (16) we eliminate P_2, place $X_1 = \frac{P_1}{Z_1}$ and substitute the second equation of (11). This gives

$$P_1 = \frac{P_3}{a - cZ_o} \tag{17}$$

corresponding to one of Professor Stewart's equations. We substitute the values of a and c given in (10) and arrive at the following equation for the exponential horn; (in this equation we have placed $x_1 = 0$ and $x_2 = L$):

$$P_1 = \frac{P_3 \epsilon^{\frac{mL}{2}}}{\cos g\,L + \frac{m}{2g} \sin g\,L + \frac{\sigma_2 K Z_o}{g\beta} \sin g\,L} \tag{18}$$

Professor Stewart derived a similar equation for the conical horn which, in connection with our equation, is discussed in the paper. All the quantities involved in this equation have been defined. The equation cannot be made use of, however, until we can calculate Z_o.

This has been done by Lord Rayleigh in his second volume of "Theory of Sound." He showed that if air is passing thru a short tube the acoustic conductivity of which is C_o, the inertia of the air gives an apparent mass factor of $\frac{\rho}{C_o}$. He also showed that if air escapes (that is, sound is radiated) from a circular opening into the outide space, the energy radiated gives a dissipational pressure coefficient of $\frac{\rho n^2}{2\pi C}$ where n is 2π times the frequency, ρ is the air density, and C is the velocity of sound. The acoustic impedance, therefore, in the sense we are using the term, of such a tube radiating sound energy into the surrounding space is:

$$Z_o = -\frac{\rho n^3}{C_o} + i\frac{\rho n^3}{2\pi C} = \rho C^2 K^2 \left(\frac{K i}{2\pi} - \frac{1}{C_o}\right) \tag{19}$$

where $K = \frac{n}{C}$.

This reason for this equation will be clear if we consider the mechanical case for vibratory displacement, corresponding to

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acoustic displacement. In mechanics we have the familiar equation

$$m\frac{d^2\xi}{dt^2}+r\frac{d\xi}{dt}+s\xi=f \tag{20}$$

If $f=Fe^{int}$ and $\xi=\xi_o e^{int}$, then

$$\xi=\frac{F}{s-mn^2+irn} \tag{21}$$

The denominator of this equation represents the mechanical impedance and if s is zero, this impedance is given by:

$$Z_m=-mn^2+irn \tag{22}$$

In the acoustic analogy, m is replaced with $\frac{\rho}{C_o}$ and the resistance coefficient, r, is replaced with $\frac{\rho n^2}{2\pi C}$. In the acoustic definition the pressure is equal to the product of the air displacement and the acoustic impedance. Hence, $\frac{\rho}{C_o}$ is the pressure per unit acceleration and $\frac{\rho n^2}{2\pi C}$ is the pressure per unit velocity. Further reference to this work may be made to Rayleigh, "Theory of Sound," volume II, pages 193-194, equation 3 (where X is unity) and page 172, equation 1. Equation (19), then, can be used to determine Z_0 which occurs in equation (18).

To study the direct and reflected waves we use equation (1) from which we get:

$$A=\frac{\begin{vmatrix}P_1 v_1\\P_2 v_2\end{vmatrix}}{\begin{vmatrix}v_1 u_1\\v_2 u_2\end{vmatrix}} \text{ and } B=\frac{\begin{vmatrix}P_1 u_1\\P_2 u_2\end{vmatrix}}{\begin{vmatrix}v_1 u_1\\v_2 u_2\end{vmatrix}} \tag{23}$$

Whence, if $x_1=0$ and $x_2=L$, and if we substitute equations (2) in (23), we get:

$$A=-P_1 \text{ and } B=-\frac{P_1\varepsilon^{-\frac{mL}{2}}\cos gL-P_2}{\varepsilon^{-\frac{mL}{2}}\sin gL} \tag{24}$$

In Appendix I we saw that:

$$A=A_1+B_1 \text{ and } B=(A_1-B_1)\,i \tag{25}$$

where A_1 is the maximum velocity potential of the direct wave and B_1 the maximum of the reflected wave. In a similar manner we can apply this relation to the maximum pressures produced by the direct and reflected waves in the pressure equation

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$$P = A_1 \epsilon^{-\frac{mL}{2}} \cos(nt - gx) + B_1 \epsilon^{-\frac{mL}{2}} \cos(nt + gx) \qquad (26)$$

which is obtained in the same manner as equation (13) in Appendix I.

From equation (25) we get

$$\frac{B_1}{A_1} = \frac{A\,i - B}{A\,i + B} \qquad (27)$$

If we substitute the results of equation (24) in (27), we get, after reduction:

$$\frac{B_1}{A_1} = \frac{-iP_1 \epsilon^{-\frac{mL}{2}} \sin gL + P_1 \epsilon^{-\frac{mL}{2}} \cos gL + P_2}{-iP_1 \epsilon^{-\frac{mL}{2}} \sin gL - P_1 \epsilon^{-\frac{mL}{2}} \cos gL + P_2} \qquad (28)$$

In this equation the square root of (-1) occurs, so that in getting absolute values of the ratio of the reflected to the direct wave it is necessary to take the vector sum of (28). By means of the above ratio, we can study the reflection effects for the exponential horn.

From equation (25) we also get:

$$A_1 = \frac{1}{2}(A - B\,i) \quad \text{and} \quad B_1 = \frac{1}{2}(A + B\,i) \qquad (29$$

In these we substitute equation (24) and get:

$$\left.\begin{aligned} A_1 &= \frac{1}{2}\left(-P_1 + \frac{P_1 i\,\epsilon^{-\frac{mL}{2}} \cos gL - P_2 i}{\epsilon^{-\frac{mL}{2}} \sin gL}\right) \\ \text{and} \\ B_1 &= \frac{1}{2}\left(-P_1 - \frac{P_1 \epsilon^{-\frac{mL}{2}} \cos gL - P_2 i}{\epsilon^{-\frac{mL}{2}} \sin gL}\right) \end{aligned}\right\} \qquad (30)$$

Using the absolute values of these equations we are able to make a study of the maximum values for the pressures in the direct and reflected waves separately.

Appendix III

The parabolic horn offers an interesting contrast with the straight pipe, the conical, the exponential, and the hyperbolic horns. In the case of the parabolic horn the area varies directly as the distance x from the small end of the horn. The velocity potential can be written as a function of x and t, according to Webster.

$$\frac{d^2\phi}{dx^2} + \frac{d \log \sigma}{dx}\frac{d\phi}{dx} = \frac{d^2\phi}{dt^2} \qquad (1)$$

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Since, in the parabola, $\sigma = \sigma_1 x$, then $\frac{d}{dx}(log\ \sigma) = \frac{1}{x}$.

Substitution in (1) gives for the parabola

$$\frac{d^2\phi}{dx^2} + \frac{1}{x}\frac{d\phi}{dx} = \frac{d^2\phi}{dt^2} \tag{2}$$

This is a familiar equation met with in vibrating diaphragms, drum heads, and the like, and has been treated in a number of places.[1] It is solved by placing $\phi = X\,T$, X being a function of x only and T being a function of t only. Substitute in (2) and we get

$$\frac{d^2T}{dt^2} + C^2\,T = 0 \tag{3}$$

and

$$\frac{d^2X}{dx^2} + \frac{1}{x}\frac{dX}{dx} + C^2\,X = 0 \tag{4}$$

The general solution of the former is

$$T = A\ cos\ K\,at + B\ sin\ K\,at \tag{5}$$

Equation (4) is the Bessel equation of zero order and its general solution is

$$X = C\,J_o(x) + D\,Y_o(x) \tag{6}$$

Consequently the general solution of (2) is

$$\phi = [C\,J_o(x) + D\,Y_o(x)]\,[A\ cos\ Kat + B\ sin\ K\,at] \tag{7}$$

The pressure is, $-\rho\frac{d\phi}{dt}$, or

$$P = -[C\,J_o(x) + B\,Y_o(x)]\,[-AKa\ sin\ K\,at + BKa\ cos\ K\,at] \tag{8}$$

and the velocity is $\frac{d\phi}{dx}$, or

$$-V = -[C\,J_1(x) + D\,Y_1(x)]\,[A\ cos\ K\,at + B\ sin\ K\,at] \tag{9}$$

On the average, the energy radiation is proportional to the product of (8) and (9), which is zero because the pressure and velocity are always 90° out of phase. The accompanying figure shows the result diagramatically. Such a case as this is unusually interesting and is highly suggestive.

It is interesting to observe that equation (8) is identical to the solution of the general equation for the vibrating drumhead or diaphragm. There are, therefore, "modes" of vibration in the infinite parabolic horn corresponding to the roots of the Bessel functions just as there are modes of vibrations for a vibrating diaphragm with which we are familiar.

In case we have a parabolic horn of finite length, we can proceed as in Appendix II and derive a number of interesting results.

[1] Rayleigh, "Theory of Sound," volume I, page 318, article 200, equation (1), where W is independent of θ. Byerly, "Fourier's Series and Spherical Harmonics," page 12, article 11. Gray and Matthews, "Treatise on Bessel Functions," chapter X.

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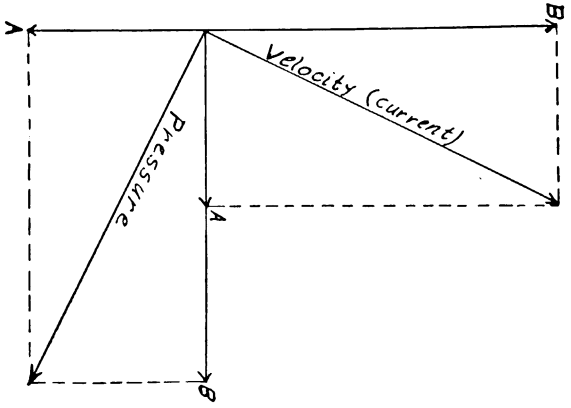
If we do this, we observe that $u = J_o(Kx)$, $u_1 = J_o(Kx_1)$, $u_2 = J_o(Kx_2)$, $v = Y_o(Kx)$, $v_1 = Y_o(Kx_1)$, $v_2 = Y_o(Kx_2)$, $u^1 = -J_1(Kx)$, $u_1^1 = -J_1(Kx_1)$, $u_2^1 = -J_1(Kx_2)$, $v^1 = -Y_1(Kx)$, $v_1^1 = -Y_1(Kx_1)$, and $v_2^1 = -Y_1(Kx_2)$.



With these results we can derive the expression for a, b, c, and d in equation (8) Appendix II. We can then derive all the equations for the parabolic horn of finite length. The resulting equations will all be of the Bessel function types and consequently the variations with frequency and distance will not be determined by the roots of the simple sine and cosine functions, but by the roots of the Bessel functions. The resonant effects, therefore, will be of a distinct type and unlike the usual ones which occur for horns of the usual curvature. Since the parabolic type of horn offers results the practical application of which require considerable imagination, it is felt that the detailed results need not be here given.

SUMMARY: (1) Experimental: By means of a vacuum tube oscillator to operate a loud speaker unit and a calibrated condenser transmitter, resistance-coupled amplifier, and a thermocouple and galvanometer, tests have been made at various frequencies to obtain the sound output from straight conical horns. The horns were made of heavy galvanized sheet iron and include lengths from 30 cm. to 305 cm. and final openings from 5 cm. to 90 cm. The results are plotted to show the sound pressure in dynes at the various frequencies up to 4,000 cycles.

Marked resonance was obtained at all the lower frequencies and the effect was still observable up to 4,000 cycles. The resonant peaks due to the horns at the lower frequencies are frequently as high as three or four times the response at the adjacent anti-resonant frequencies. The ratio approaches unity for all horns at the higher frequency range and for all the horns with the large openings there are not large resonant peaks. At the peaks the resonant is most marked for the horns with small openings, and the "valleys" are not "filled in." The larger solid angle, however, causes sufficient resonance at all frequencies above the fundamental to fill in the valleys and to cut off the sharpness of the peaks. The result is a more uniform sound output. For the

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horns of each length and varying solid angle there is a particular solid angle and opening which give the optimum sound output. The best results were obtained with the longer horns of not too small or too large an opening.

Impedance analysis indicated marked effects of the horns on the diaphragm near the fundamental frequency of the diaphragm.

Dissipation due to sound radiation from the horns near the natural frequency of the diaphragm is about equivalent to the dissipation of the natural resistance of the diaphragm itself. This is not the case if the horn is not a good radiator of sound.

(2) Theoretical: Based on the assumption of plane waves, equations for pressure, velocity, sound radiation, and phase of the unlimited straight pipe and exponential and parabolic horns are reviewed or developed. The corresponding equations for the straight conical horns with the source at the vertex are reviewed and a comparison made among them. In the case of the parabolic horn the pressure and velocity are always out of phase and no sound energy is radiated. The velocity and pressure equations of a parabolic horn have solutions similar to a vibrating membrane and therefore the series of resonant and anti-resonant frequencies is determined by the maxima and minima of Bessel functions rather than by the maxima and minima of sine or cosine functions as in the pipe and exponential horn.

The equations for pressure and velocity show these equations to be in phase at all frequencies for the pipe; and sound radiation along it is independent of frequency. In the case of the cone with the source at the vertex and the exponential horn, these quantities are in phase for all the high frequencies and for this frequency region the sound radiation along the cone exceeds that along the exponential horn. Over the intermediate or middle frequency region the sound radiation along the exponential horn exceeds that along the cone. At the low frequency region the radiation along the cone again exceeds that along the exponential horn. Experimentally, these statements have been shown to hold at the low and intermediate frequencies for horns of finite initial openings. In this case, however, the horns are much alike at the high frequencies.

There is a finite frequency where the power radiation becomes zero and the velocity and pressure become 90 degrees out of phase for the exponential horn whereas this does not occur in the case of the cone until zero frequency is reached.

In the case of horns of finite lengths, Webster's method of treatment is extended to cover in detail the exponential horn and a comparison is made between it and the straight pipe and cone. The acoustic impedance of the large end is studied in detail and an equation is derived for the exponential horn to show the sound pressure just outside the large end in terms of the pressure at the receiver diaphragm. This equation shows all the phenomenon of resonance of exponential horns and can be compared with a similar equation for the conical horn derived by Stewart.

The solution of the parabolic horn of finite length has been worked out and briefly indicated.

Equations for the direct and reflected sound energy from the open end have been derived and stated in brief form. By means of these we are able to study end reflection in conjunction with the resonance phenomenon which is of great importance in our use of loud speaker horns as amplifiers of sound. In addition to resonance, the horn at the small end has a high acoustic impedance which gradually diminishes to a small value at the large end. In this way the horn greatly increases the sound radiation from a receiver diaphragm because it adjusts the acoustic impedance of the small end to match the unit and in the same way it adjusts the impedance at the large end to fit the surrounding air.

This study has led to a comprehensive understanding of the performance of loud speaker horns and will undoubtedly lead to new and improved types of loud speakers.

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THE LIMIT OF REGENERATION*

By

N. C. Little

(Bowdoin College, Brunswick, Maine)

It is a well-known fact that the regenerative effect of a vacuum tube on an oscillatory circuit may be considered as the reduction of the effective resistance of that circuit. Further, if the regeneration is pushed too far, that is, if the coupling between grid and plate circuits is increased beyond a certain critical value, the effective resistance becomes negative and the system gives rise to self-sustained oscillations of a frequency corresponding to zero reactance. In the use of a regenerative system for radio reception, the borderland region between the purely regenerative state and that of spontaneous oscillation is of paramount importance. At the suggestion of Professor E. L. Chaffee, research was undertaken at Cruft Laboratory, Harvard University, to investigate the limit of pure regeneration and the conditions of stability at that limit.

The system studied consisted of a simple oscillatory circuit with the regeneration produced by the inductive coupling of a reaction coil in the plate circuit of the tube. The adjustment of this reaction coupling was made extremely fine. In fact, by employing two reaction coils, one of a comparatively large number of turns and one of a single turn, it was possible to vary the coupling by steps of one part in a million. By this means, it was possible to arrive at that critical adjustment just beyond which the system would give rise to self-sustained oscillations. This adjustment, however, was obtained only by extremely patient manipulation, first by increasing the reaction coupling in minute steps and then, after each increase, ascertaining that the system was not oscillating of itself. However, when this critical adjustment is obtained, the system has zero reactance and, as will be shown later, practically zero resistance to the frequency which it is desired to receive. The state of regeneration, corresponding to this critical adjustment, is the practical limit to which pure

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regeneration may be pushed in the laboratory and is called *critical regeneration.* In its other details the system studied was not unique. The grid and filament of the tube were connected, as usual, across the condenser of the oscillatory circuit. The load in the plate circuit was made as small as possible.

The preliminary work was done at a frequency of 1,000 cycles. By balancing, thru a series of mutual inductances, the electromotive force impressed upon the oscillatory circuit with that due to the alternating component of the plate current, a measure of the relative magnitudes of the impressed and received signals was obtained. Measurements were made at that limit of regeneration just beyond which the system gives rise to self-sustained oscillations, namely, at critical regeneration. Results obtained with VT-1 and a UV-201, both used with varying amounts of grid bias, show that *the relative magnitude of received signal in the plate circuit to that impressed on the grid is inversely proportional to the latter.* This is called the inverse signal strength law. It states that the response of a system adjusted to critical regeneration is independent of the strength of the impressed signal, that no matter how weak the oscillating field surrounding an antenna may be, if the regeneration is pushed to its limit, a finite signal may always be obtained.

It should be remembered, however, that this law applies only to the state of critical regeneration, a condition rarely obtained in engineering practice and only in the laboratory with extremely cautious adjustment. The effective resistance of the regenerative system under these circumstances, however, may be made so low that the temperature variations of the resistance in the inductances are appreciable in comparison. Transient phenomena, which under ordinary circumstances may be neglected, now may readily be observed, as the sound in the telephone receivers is heard to die away gradually upon the removal of the impressed signal. In fact, assuming the original sound in the telephone receivers to be a million times that of the sound just audible and also proportional to the square of the current in the oscillatory circuit, it is possible from the exponential law of current decay, to estimate the effective resistance of the system. At the lower frequency (1,000 cycles), where the inductance was of the order of 300 millihenrys with an initial resistance of 25 ohms, the effective resistance at critical regeneration was found to be 0.014 ohms. At the higher frequency (750 kilocycles), where the inductance was of the order of 0.2 millihenrys and the oscillatory circuit had an initial high frequency resistance from 1 to 10 ohms,

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this critical effective resistance was found to be only 0.00014 ohms.

The effective resistance measured by the above method is that which obtains when the potential impressed on the grid is very small. In order to measure the effective resistance when that potential is large, it is necessary to study the form of the resonance curve. For this purpose a radio-frequency carrier-wave completely modulated by a single audio frequency is used. It may be shown, both graphically and analytically, that for the resolution of a single band, that is, for the resonance curve of the regenerative system to show in addition to a central maximum two adjacent side maxima, it is necessary that the damping constant, $R/2L$, be less than 0.06 of the modulating frequency. In the resonance curves obtained, the side bands were easily detected. The corresponding effective resistance, however, was of the order of an ohm. Thus, it is seen that the effective resistance of a system using extreme values of regeneration is not a constant but a function of the potential applied to the grid.

That such must be the case is apparent at once when one considers that the effective resistance depends on the tube constants. Consider, for example, the internal plate-filament resistance. Characteristics of this, plotted against grid potential, consist of a series of curves concave upward with their maximum points moving toward the values of negative grid potential as the filament current or plate voltage is increased. It is at once apparent that the average or effective values of this tube constant, which obtains when the operating point moves over a finite portion of the arc of this characteristic, is greater than the static value corresponding to the mid-point of this arc. On the other hand, for the amplification factor with a characteristic concave downward, the effective operating value for large oscillations of the grid potential is less than that for small. The effect, then, of an increase in the potential variation applied to the grid is to increase the plate-filament resistance and to decrease the amplification factor of the tube of the regenerative system. These two changes result in an increase of the effective resistance of the system. Thus the discrepancy between the two values of the effective resistance is due to the non-linear characteristics of the tube.

Considering as a first approximation the characteristic curves to be of the second degree, it may be shown that the relation between current amplitudes and effective resistance is such as to give the inverse signal strength law mentioned above, and that

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the condition of critical regeneration is essentially stable, that is, an impressed signal will not set the system into a state of self-sustained oscillations.

In conclusion, the results of this study of regeneration give the following suggestions for the use of the regenerative method in practical cases: First, the reaction coupling as employed on the majority of commercial receiving sets does not permit of sufficiently fine adjustment. Such sets should be equipped with a vernier coupling coil of a single turn. By this means, the apparent instability due to overstepping the state of critical regeneration on account of the coarseness of the coupling adjustment, will be eliminated.

Second, the inverse signal strength law shows that the weaker signals will be as effective as the stronger ones in producing a response in the telephones. It is also well-known that a circuit with little resistance has a very sharp resonance curve. Thus, the regenerative method is ideal for selective radio-telegraph reception. For radio-telephony, however, the variation with signal strength will produce distortion and the peak of the resonance curve may be too narrow to include the complete band of frequencies it is desired to receive. For this reason, probably the extreme regeneration used in this investigation is not suited to commercial radio telephony. However, this research does suggest that better results will be obtained if the tube used for regenerating is not also used for rectifying. The rectifying properties of a tube depend upon the curve of its characteristic, and it is this very curvature which introduces the variation with signal strength. Thus, using a crystal rectifier in conjunction with a regenerating tube, operated on the linear portion of its characteristic and with the adjustment slightly below that of the state of critical regeneration, one should have a simple efficient method for radio-telephonic reception.

Finally, in regard to the sensitiveness of the method. A system with zero resistance should give an infinitely large response to its resonant frequency. Or, stating the fact in terms of radio-reception, a circuit adjusted for critical regeneration will give a finite sound in the telephones for an infinitely weak signal. It makes no difference to the regeneration whether the resistance for which it is compensating is due to the losses in the wire of the inductances and in the dielectric of the capacities, or to the energy which is re-radiated. In any case, at critical regeneration, the system is in a stable state with a net effective resistance, to an incoming signal of the proper frequency, of practically zero. Thus, other

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than that necessitated by the difficulty in arriving at the adjustment for critical regeneration, there is no limit to the sensitiveness of the simple regenerative method.

SUMMARY: It is shown that, at the limit of regeneration, the relative magnitude of received signal in the plate circuit to that impressed on the grid is inversely proportional to the latter. This inverse signal strength law is explained in terms of the characteristics of the tube. Two methods of determining the effective resistance of a regenerative system are outlined. Practical suggestions are offered for the design of regenerative systems.

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ON THE CALCULATION OF THE INDUCTANCES AND CAPACITIES FOR A MULTI-RANGE OR OTHER CONSECUTIVE SERIES OF TUNED TRANSMITTING OR RECEIVING CIRCUITS, THE TOTAL RANGE AND ACCURACY REQUIRED BEING GIVEN*

By

J. Erskine-Murray

(His Majesty's Signal School, Royal Naval Barracks, Portsmouth, England)

1. When a large range of natural frequencies (or wave lengths) has to be covered by one instrument, it is usual to use a number of capacities or inductances of fixed values in combination with a single variable capacity or inductance of comparatively small magnitude. The total range is thus divided into a series of consecutive scales and at the commencement of each scale a new fixed inductance or capacity, or both, is switched in and the scale is then covered by use of the variable. The values of these quantities have in general been arrived at by a process of "trial and error" and the possible accuracy of reading has varied from scale to scale.

In order to place the design of such an instrument on an engineering basis, the following method of calculation has been worked out. By its aid the design of a multi-range wavemeter, transmitter or receiver, to fulfil given requirements of range and accuracy, becomes a simple and straightforward matter.

2. Symbols used in this paper:

λ = wave length in meters.

$\lambda_1 . . \lambda_n$ = wave lengths at commencements of intermediate ranges (or scales).

f = frequency in cycles per second.

$f_1 . . f_n$ = frequencies at commencements of intermediate ranges.

L = inductance in microhenrys (mics.).

$L_1 . . L_n$ = inductances, fixed, at commencements of scales.

L_v = maximum value of variable inductance.

$L_{v\,o}$ = minimum value of variable inductance.

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L_w = portion of L_t required to cover scale without overlapping next scale.

L_F = actual fixed inductance $(L_F)_n = L_n - L_{to}$.

C = capacity in milli-microfarads (mifs.).

C_n and so on = C with subscripts similar to L_n and so on.

$d°$ = smallest angular difference readable on scale of variable.

c = percentage accuracy, of λ or f.

b = minimum frequency difference, in cycles, to be readable.

a = a percentage of variable; $a/100$ = fraction of scale of variable used to cover a range without overlapping the next.

$(LC)_n$ = oscillation constant for commencement of nth scale.

3. Total Range—Let the total range of wave lengths to be covered be from λ_1 to λ_m, the corresponding frequencies being f_1 to f_m and LC values $(LC)_1$, to $(LC)_m$. $(LC)_m$ is thus the maximum value on the scale commencing at $(LC)_{m-1}$, and there are only $(m-1)$ scales required.

4. Accuracy—Let the desired accuracy of wave length reading thruout the nth intermediate range be c_n percent of the shortest wave length λ_n of this range. This will ensure that the accuracy thruout the range λ_n to λ_{n+1}/λ is not less than c percent of the wave length read. The cases where accuracy is defined as a percentage of frequency, or as a difference of frequency, are considered in (6) and (7) below.

5. The possible accuracy depends on the ratio, λ_{n+1}/λ_n, of the wave lengths at the beginning and end of the scale, and on $d°$, the minimum angular difference readable with certainty on the scale of the variable condenser or inductance. In the case of a receiving circuit or wavemeter, $d°$ depends not only on the radius and graduation of the scale, but also on the damping of the circuit and type of resonance current indicator used.

6. Thus, in order to obtain the desired accuracy, we find that for a movement of $d°$ on the scale of the variable, λ_n may increase to $\lambda_n\left((1+\frac{c}{100}\right)$. In the whole scale of the variable, say 0° to 180°, there are $180/d$ such steps, assuming that the calibration curve of the variable is substantially a straight line. Each step represents an equal increase of wave length (see (4)); the accuracy therefore increases towards the upper ends of the scale. Hence, putting in the whole variable increases λ_n to

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$$\lambda_n\left(1+\frac{180}{d}\frac{c}{100}\right).$$

Thus if the whole variable were used, we should have the end of the scale and commencement of the next at

$$\lambda_{n+1}=\lambda_n\left(1+\frac{1.8\,c}{d}\right).$$

However, for practical reasons it is necessary that successive scales should overlap. Let us therefore take only a percentage a (say 80 percent) of the variable scale as giving the whole intermediate range. Hence to provide for the overlap we find

$$\lambda_{n+1}=\lambda_n\left(1+\frac{180}{d}\frac{c\,a}{10{,}000}\right) \qquad (A)$$

and when the variable is being moved from $\frac{a}{100}\times 180°$ to $180°$ the scale will be overlapping the next in order.

We have, therefore, found that to obtain the desired accuracy of reading, the intermediate range commencing at λ_n must finish at the value of λ_{n+1} given in (A); and since $(L\,C)$ values vary as the square of the wave length

$$(L\,C)_{n+1}=(L\,C)_n\left(1+\frac{180\,c\,a}{10{,}000\,d}\right)^2 \qquad (B)$$

7. If frequencies be reckoned instead of wave lengths

$$f_{n+1}=f_n\left(1-\frac{180\,c\,a}{10{,}000\,d}\right) \qquad (C)$$

Since the frequency decreases for increasing capacity, and

$$\frac{(L\,C)_{n+1}}{(L\,C)_n}=\frac{(1/f_{n+1})^2}{(1/f_n)^2}=\frac{f_n{}^2}{f_{n+1}{}^2}$$

Hence

$$\frac{(L\,C)_{n+1}}{(L\,C)_n}=\frac{1}{\left(1-\frac{180\,c\,a}{10{,}000\,d}\right)^2} \qquad (D)$$

and the limits of successive scales may be determined as for wave lengths.

8. If the accuracy be defined as an actual difference of frequency, b cycles (for instance 100 cycles), instead of a percentage difference, a system which is convenient when a heterodyne receiver is used, we have

$$f_{n+d^\circ}=f_n-b$$

therefore

$$f_{n+1}=\left(f_n-b\frac{180}{100}\frac{a}{d}\right) \qquad (E)$$

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and $$\frac{(L\,C)_{n+1}}{(L\,C)_n}=\frac{f_n^2}{f_{n+1}^2}=\frac{f_n^2}{\left(f_n-b\,\frac{180\,a}{100\,d}\right)^2} \qquad \text{(F)}$$

So that in this case also the limit of each intermediate range can be found when the total range and accuracy have been given.

9. Since a variable condenser cannot be got to ge down to zero capacity, but has some lower limit C_{vo}, its rango is actually C_{vo} to C_v, if C_v be its maximum. C_{vo} is thus a fixed capacity which forms part of the calculated fixed capacity for the range. Thus the magnitude of the fixed condenser to be switched in at the beginning of the nth range is $C_F=C_n-C_{vo}$, where C_n is the calculated fixed capacity. In general, if C_v be chosen of any convenient dimensions, for instance 1 milli-microfarad, the fixed capacity and inductance to fulfil the desired conditions are then determinable as is shown in sections (10), (11), (12), and (13).

10. Determination of Fixed Capacities and Inductances, the Variable Being a Capacity—Let L_n and C_n be the fixed inductance and capacity for the range beginning at λ_n or $(L\,C)_n$. This may, for instance, be the first scale of the total range required, in which case λ_n is known. In other cases λ_n is determinable by working out the previous ranges starting with λ_1, or by the method of (23).

Then for the short-wave end of this scale

$$(L\,C)_n=L_n\,C_n$$

and at the other limit where the variable capacity

$$=C_w=(C_v-C_{vo})\,\frac{a}{100}$$

$$(L\,C)_{n+1}=L_n\,(C_n+C_w)$$

See (6).

Hence $$\frac{(L\,C)_{n+1}}{(L\,C)_n}=\frac{C_n+C_w}{C_n}$$

in which C_w and the $(L\,C)$ values are known, gives C_n. The cases for different conditions of accuracy are given below.

11. If the accuracy be a percentage of wave length, as in (6)

$$\frac{C_n+C_w}{C_n}=\frac{(L\,C)_{n+1}}{(L\,C)_n}=\frac{\lambda_{n+1}^2}{\lambda_n^2}=\left(1+\frac{180\,c\,a}{10{,}000\,d}\right)^2$$

Hence $$C_n=\frac{C_w}{\left(1+\frac{180\,c\,a}{10{,}000\,d}\right)^2-1}=\frac{C_w}{\frac{1.8\,c\,a}{100\,d}\left(2+\frac{1.8\,c\,a}{100\,d}\right)} \qquad \text{(G)}$$

and since c, a, and d are known, this gives C_n and the fixed con-

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denser required is $C_n - C_{vo}$. Also from $(LC)_n = L_n C_n$, we get L_n the fixed inductance.

12. If the accuracy be defined by percentage of frequency we get from (7) by a process similar to that of (11)

$$C_n = C_w \frac{\left(1 - \frac{180\,c\,a}{10{,}000\,d}\right)^2}{\frac{1.8\,c\,a}{100\,d}\left(2 - \frac{1.8\,c\,a}{100\,d}\right)} \qquad (H)$$

and L_n is calculable from $(LC)_n$ as in (11).

(13) If the accuracy be defined as a given change in number of cycles, $b\sim$, we find from (8)

$$\frac{(LC)_{n+1}}{(LC)_n} = \frac{{f_n}^2}{\left(f_n - b\frac{180\,a}{100\,d}\right)^2}$$

Hence

$$C_n = C_w \frac{\left(f_n - b\frac{180\,a}{100\,d}\right)^2}{b\frac{180\,a}{100\,d}\left(2f_n - b\frac{180\,a}{100\,d}\right)} \qquad (I)$$

and for this case also the fixed capacity and inductance for each intermediate range are easily determinable. Examples of the calculation of several multi-range circuits are given in (16) and following.

14. Instead of using the equations (G), (H), or (I) to determine the fixed capacities and inductances for the various intermediate ranges, we may employ a graphical method. Thus a series of hyperbolas may be drawn in which L and C are coordinates, one for each value of $(LC)_n$ as determined in (6), (7), and (8), starting with the required initial (LC) of the shortest wave to be received or produced in the instrument. By inspection of each successive pair of curves (see Figure 1), a point P may be found on the curve of lesser (LC), such that C_w, measured parallel to the axis of C, will just reach the next curve at some point Q. The inductance L_1 and capacity C_1 of the point P are the fixed values required in order that the variable C_w may cover the first section of the total range with the desired accuracy. By repeating this process from curve to curve the fixed capacity and inductance for each range is obtained.

15. In certain circuits it is necessary that the ratio L/C should be greater or less than a certain limiting value, k, where $k = \frac{L}{C} = \frac{R^2}{4}$; for instance, in order to control the damping of the

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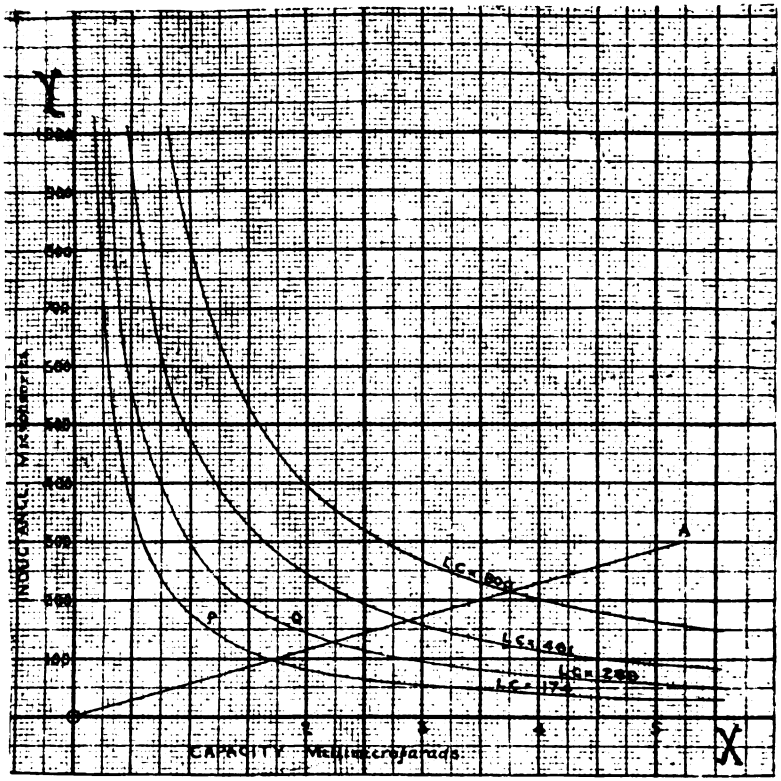

FIGURE 1

circuit and thus facilitate or hinder the production of oscillations. The graphical method of (14) lends itself readily to the introduction of such a condition. Thus if a line $O\,A$ be drawn from the origin of co-ordinates making an angle the tangent of which is k with the axis of capacity, each hyperbola will be cut in two by $O\,A$. If L/C is to be greater than k, only that part of the diagram between $O\,A$ and $O\,Y$ may be used, since for any point in this area $L/C > k$. If L/C is to be less than k the portion between $O\,A$ and $O\,X$ must likewise be used. In cases like these it will generally be convenient to defer the choice of the variable condenser or inductance until the curves have been examined.

16. THE VARIABLE AN INDUCTANCE—If a variable inductance be used instead of a capacity, the process of finding the values of fixed capacities and inductances can be carried out from the same curves by working on the other co-ordinate. The values of $(L\,C)_n$ are the same as in (14), since they depend only on c, a, and d. (See (6)). The algebraic method is also similar, and

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if L_v, L_{vo} and L_w be the maximum, minimum and reduced maximum (see (6)), of the variable inductance, we obtain in place of the equations (G), (H), and (I), the values of the fixed inductance in terms of the variable as

$$L_n = \frac{L_w}{\frac{1.8\,c\,a}{100\,d}\left(2+\frac{1.8\,c\,a}{100\,d}\right)} \qquad (G')$$

for percentage wave length accuracy, and

$$L_n = \frac{L_w\left(1-\frac{1.8\,c\,a}{100\,d}\right)^2}{\frac{1.8\,c\,a}{100\,d}\left(2-\frac{1.8\,c\,a}{100\,d}\right)} \qquad (H')$$

for percentage frequency accuracy, and

$$L_n = L_w\frac{\left(f_n-b\frac{180\,a}{100\,d}\right)^2}{b\frac{180\,a}{100\,d}\left(2f_n-b\frac{180\,a}{100\,d}\right)} \qquad (I')$$

for a fixed frequency difference as accuracy.

17. EXAMPLE—As an example of the method let us assume that a series of circuits is required in accordance with the following data:

(I) Total range of wave lengths 800 m. to 2,000 m.

(II) Accuracy to be to 0.1 percent of λ, that is, 1 m. in 1,000 m.

(III) Smallest readable angle in scale $d° = 0°.5$.

(IV) Variable condenser $C_{vo} = 0.04$ mif. to $C_v = 1.06$ mif.

(V) $C_w = (C_v - C_{vo})\frac{80}{100} = 0.816$ mif.

From the condition (II) the limits of the intermediate ranges are determinable thus:

$$(LC)_2 = (LC)_1\left(1+\frac{1.8\,c\,a}{100\,d}\right)^2 = (LC)_1\,(1+0.288)^2 = (LC)_1\,(1.66)$$

To allow a margin let us start at 785 m. instead of 800 m., then

$(LC)_1 = 173.6$ mics. $\times$ mifs.

$(LC)_2 = 289.1$ mics. $\times$ mifs.

$(LC)_3 = 480.0$ mics. $\times$ mifs.

$(LC)_4 = 796.8$ mics. $\times$ mifs.

Now

$$C_1 = \frac{C_w}{0.288\,(2.288)} = \frac{0.816}{0.660} = 1.236 \text{ mifs from } (G),$$

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therefore $C_F = C_1 - C_{vo} = 1.236 - 0.04 = 1.196$ mifs. and also $L_1 = 173.6/1.236 = 140.4$ mics. The others are as shown in the table. It should be noted that, as the accuracy is to be the same thruout, C_F has the same value for all ranges, as may be seen from equation (G). Table I shows the results of calculation for the example. The overlap of each pair of scales is found by taking $L_n\,(C_n + C_v - C_{vo})$ as the largest $(L\,C)$ of the scale.

Thus there will be four scales required to cover the range under the conditions stated. One fixed capacity will serve for all ranges and the inductance for each range is shown in the table.

TABLE I

Scale Number	$(L\,C)_n$ mic. × mif.	λ meters	C_n mifs.	L_n mics.	Other Numerics
1	173.6	785 – 1061	1.236	140.4	$C_F = 1.236 - 0.04 = 1.196$ mif.
2	289.1	1013 – 1360	1.236	234.0	$C_v = 1.06$ $C_F + C_v = 2.256$
3	480.0	1310 – 1763	1.236	388.5	$C_{vo} = 0.04$ $C_v = 0.816$
4	796.8	1686 – 2282	1.236	645.0	$d° = 0.5$ $a = 80\%$ $c = 0.1\%$

18. If a variable inductance, instead of a variable condenser, be used in the instrument calculated in Table I, we find the same $(L\,C)$ values for the limits of the intermediate ranges of the calibration curve of the variable inductance is approximately straight. For the values of L_n we have the equation (G') of (15)

$$L_n = \frac{L_w}{\frac{1.8\,c\,a}{100\,d}\left(2 + \frac{1.8\,c\,a}{100\,d}\right)}$$

where, as before $c = 0.1$ percent, $a = 80$ percent, $d = 0°.5$. Hence, if, for instance, $L_v = 140$ mics., and $L_{vo} = 15$ mics., then

$$L_w = \frac{80}{100}\left(L_v - L_{vo}\right) = 100 \text{ mics.},$$

from G',

$$L_1 = \frac{100}{0{,}288\,(2.288)} = 156 \text{ mics.}$$

and since the accuracy is to be the same thruout the whole range of the instrument $L_2 = L_1 = \ldots$, and from the values of $(L\,C)$ calculated as for Table I, we find C_1, C_2, and so on.

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TABLE II

Scale Number	$(L\,C)$ mic.×mif.	λ meters	L_n mics.	C_n mifs.	Other Numerics
1	173.6	785−1054	156.0	1.112	$L_r = L_n - L_w$ $= 156 - 15$ $= 141$ mics.
2	289.1	1013−1362	156.0	1.852	$L_w = 100$ mics. $L_r + L_w = 281$ mics.
3	480.0	1310−1750	156.0	3.078	$d° = 0°.5$ $c = 0.1\%$
4	796.8	1686−2267	156.0	5.110	$a = 80\%$

19. Similar calculations are as easily made for accuracy based on percentage of frequency by use of the equations (C), (D), (H), and (H').

20. The calculation for an accuracy based on fixed frequency difference gives interesting results and illustrates how the difficulty of heterodyning at high frequencies may be overcome, at least theoretically, by the proper choice of ranges.

Let us assume the same total wave length range as before, that is, 800 m. to 2,000 m.—hence

(I) Total range of frequencies 375,000−150,000 cycles.

(II) Accuracy to be defined as a change of frequency of 100 cycles per 0°.5 thruout.

(III) Variable condenser 0.04 to 1.06 mifs. and $C_w = 0.816$ mifs.

Using equation (E) the scales come out as shown in Table III, the first scale commencing at 382,500 cycles in order to give a margin for constructional imperfections of the instrument. The overlap between successive scales is about 7,000 cycles in each case. The calculations are as follows, from (8), equation (E),

$$f_{n+1} = \left(f_n - 100\,\frac{180}{0.5}\,\frac{80}{100}\right) = (f_n - 28{,}800)$$

Commencing with scale (1), $f_1 = 382{,}500$ corresponding to 784.3 m.

$$f_2 = 382{,}500 - 28{,}800 = 353{,}700$$

and as the accuracy is to be the same thruout the total range, the difference is common and each f_n is found by subtracting 28,800 cycles from the one before. The Table III shows the results. From equation (I)

$$C_n = C_w \frac{(f_n - 28{,}800)^2}{28{,}800\,(2f_n - 28{,}800)}$$

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Thus

$$C_1 = 0.816 \frac{(353{,}700)^2}{28{,}800\,(2\times 382{,}500 - 28{,}800)}$$

$$= 4.91 \text{ mifs.}$$

But $(L\ C)_1 = 174.2$, therefore $L_1 = 35.2$ mics., and the other L's and C's are similarly calculated. The overlaps of the scales are, as before, found by taking C_p instead of C_w to get the largest $(L\ C)$ value of each range.

TABLE III

Scale Numbers	Frequency Cycles per second	$(L\ C)_n$ mic.× mif.	C_n mifs.	L_n mics.	C_F mifs.	Other Numerics
1	382,500 −347,000	173.2	4.91	35.2	4.87	$C_p = 1.06$
2	353,700 −318,000	202.6	4.43	45.7	4.39	$C_w = 0.816$
3	324,900 −288,500	240.1	4.01	59.9	3.97	$C_{ss} = 0.04$ $d° = 0°.5$
4	296,100 −255,900	289.1	3.58	80.8	3.54	$b = 100\,\omega$ $a = 80\%$
5	267,300 −231,500	354.7	3.17	111.9	3.13	
6	238,500 −203,200	445.6	2.78	100.4	2.74	
7	209,700 −173,300	576.4	2.36	244.0	2.32	
8	180,900 −145,000	774.5	1.95	397.0	1.91	

It should be noted that in order to obtain the same heterodyne accuracy thruout it is necessary to change both fixed capacity and fixed inductance as one advances from one scale to the next.

21. If a variable inductance be used, the total range being the same, the calculations are similar to those of (20), but are based on equations (E) and (I') and give the results shown in Table IV. The intermediate ranges are the same as in Table III since they depend on the accuracy only. With the numerics given in the table the calculation of L_n takes the form

$$L_n = 100 \frac{(f_{n+1})^2}{28{,}800\,(2\times f_n - 28{,}800)}$$

and it is advisable to get these values correct to four figures, by using seven figure logarithms or a Fuller slide rule.

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TABLE IV

Total Range, 375,000 ω to 150,000 ω = 800 m. to 2,000 m.

Scale Numbers	Frequency ω	$(L\,C)_n$ mic.× mifs.	C_n mifs.	L_n mics.	L_P $L-L_{\nu}$	Other Numerics
1	382,500 −346,500	173.2	0.319	543	528	$L_{\nu}=100$ mics.
2	353,700 −317,000	202.6	0.408	497	482	$L_{\nu}=$ 15 mics.
3	324,900 −287,500	240.1	0.532	451	436	$L_{\nu}=140$ mics.
4	296,100 −260,500	289.1	0.715	405	390	$a=80\%$
5	267,300 −230,000	354.7	0.988	359	344	$b=100\,\omega$
6	238,500 −202,000	445.6	1.420	314	299	$d=0°.5$
7	209,700 −173,000	576.4	2.148	268	253	
8	180,900 −144,500	774.5	3.485	222	207	

With these values a difference of 0°.5 on the scale of the variable gives 100 ω difference of frequency.

22. It may be noted that a useful check on the figures for C_n in Table III and L_n in Table IV may be obtained by plotting them as a curve with frequency as abscissa. In both cases the curve should be a straight line and any arithmetical error is shown up at once.

23. It is possible to find the number of scales required and the initial wave length of each, directly from the original data without successive calculation of each range. Thus, if the accuracy be uniform thruout, let

$$\left(1+\frac{180\,a\,c}{10{,}000\,d}\right)=k \qquad \text{as in (6) } (A),$$

therefore

$$\lambda_n=\lambda_1\,(k)^{n-1} \qquad (K)$$

$$(L\,C)_n=(L\,C)_1\,(k)^{2(n-1)} \qquad (L)$$

Hence from (K)

$$log\,\lambda_n=log\,\lambda_1+(n-1)\,log\,k$$

thus if λ_n and λ_1 be the given limits of the total range and if k be

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the same for all ranges, $(n-1)$, the number of scales, is determinable. More generally, if k vary from scale to scale

$$\lambda_n = \lambda_1 (k_1 \times k_2 \times k_3 \cdot \cdot \cdot)$$

and $$\log \lambda_n = \log \lambda_1 + \log k_1 + \log k_2 + \text{ and so on,}$$

hence λ_n may be determined directly from the conditions of accuracy. The same is true for $(L\,C)_n$, since

$$(L\,C)_n = (L\,C)_1 (k_1 \times k_2 \times \cdot \cdot \cdot)^2$$

$$\log (L\,C)_n = \log (L\,C)_1 + 2 (\log k_1 + \log k_2 \cdot \cdot \cdot)$$

If the accuracy be defined by a difference of frequency as in (8) we have

$$(L\,C)_n = (L\,C)_1 \left[\frac{f_1^2}{f_n^2}\right] = (L\,C)_1 \left[\frac{f_1^2}{(f_1 - (n-1)\,\Delta)^2}\right]$$

where $\Delta = b \dfrac{180\,a}{100\,d}$ = difference in frequency between the initial frequencies of two successive ranges.

Thus the formulas of this section form an alternative method for the calculation of the limits of the intermediate ranges.

SUMMARY: It is often necessary to use a series of consecutive tunablr circuits to cover a large range of wave lengths or frequencies. The papee describes a method of determining the constants of each circuit when the total wave length range and the accuracy of setting required in each intermediate range are given. Overlap between successive scales is provided for, and a method is indicated by which further conditions, such for instance as a limiting value of the damping of each circuit, may be introduced. The paper thus reduces the calculations of such circuits from an empirical problem to a straightforward engineering proposition.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

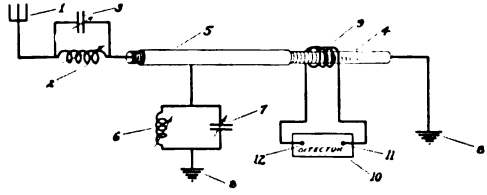
ISSUED MAY 6, 1924–JUNE 17, 1924

BY

JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, D. C.)

1,493,024—Louis Cohen and Joseph O. Mauborgne, filed August 6, 1920, issued May 6, 1924.



NUMBER 1,493,024—Radio Signaling

RADIO SIGNALING, utilizing a wave coil for the reduction of static disturbances and other interferences in the reception of radio signals. A long helix uniformly wound with wire is connected thru a tuning circuit with the antenna and to ground. The radio receiving apparatus is connected with the wave coil and an adjustable grounded metallic tube is provided in relation to the wave coil, which tube is electrically connected in a tuned circuit with ground. The adjustment of the metallic tube and the grounded loop circuit operates to change the electrical characteristics of the coil. Persistent waves set up in the coil will be received while foreign electrical disturbances which may act on the antenna, such as electrostatic effects causing a shock excitation being of a steep wave front are rapidly damped out in their travel thru the wave coil.

1,493,148—Montgomery B. Cohen, filed July 23, 1920, issued May 6, 1924.

*Received by the Editor, July 18, 1924.

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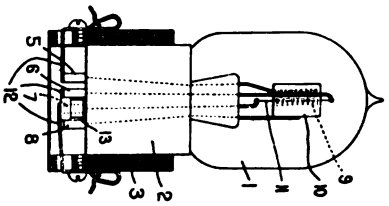
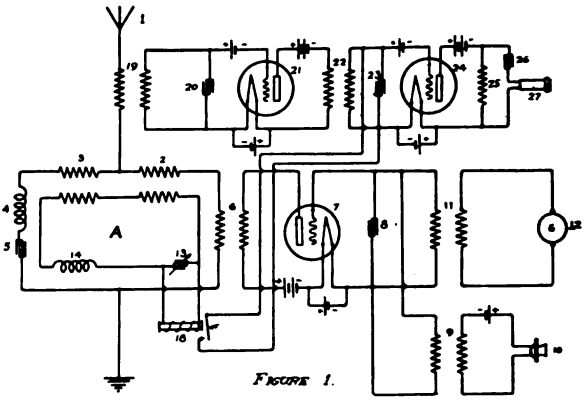


NUMBER 1,493,148—Fuse for Vacuum Bulbs

FUSE FOR VACUUM BULBS for protecting the filament from destructive effects of excessive current. The fuse consists of a cylindrical member which is applied to one of the bulb terminals and constructed to engage the corresponding socket contact.

1,493,151—Mihran M. Dolmage, filed September 17, 1920, issued May 6, 1924.



NUMBER 1,493,151—Radio Duplex Signaling System

RADIO DUPLEX SIGNALING SYSTEM which does not require the manual operation of switches to transfer the system from transmitting or receiving or vice versa. An auxiliary circuit tuned to the currents of the transmitting frequency is provided in association with the transmitting antenna. A control relay is connected in the auxiliary circuit in such manner that it is operated when currents exist in the auxiliary circuit. The operation of the relay causes the receiving apparatus to be protected against the transmitting currents during the periods of transmission.

1,493,696—Stanley R. Mullard, filed November 9, 1920, issued May 13, 1924. Assigned to the Mullard Radio Valve Company, London, England.

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THERMIONIC OR OSCILLATION VALVE, wherein a cylindrical plate electrode is supported within the tube by a collar member which surrounds a re-entrant portion of the tube which extends into one end of the tube.

1,494,347—Lloyd Espenschied, filed April 30, 1920, issued May 20, 1924. Assigned to American Telephone and Telegraph Company, Incorporated, New York.

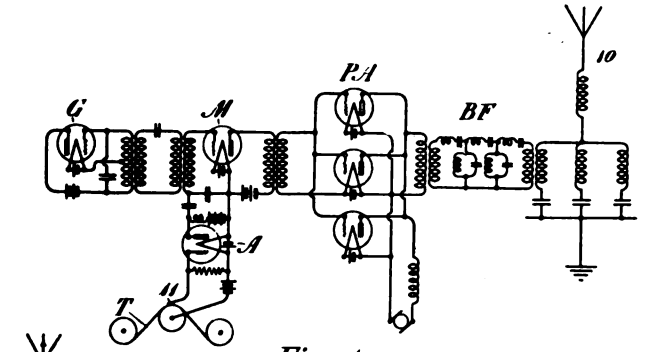

Fig. 1

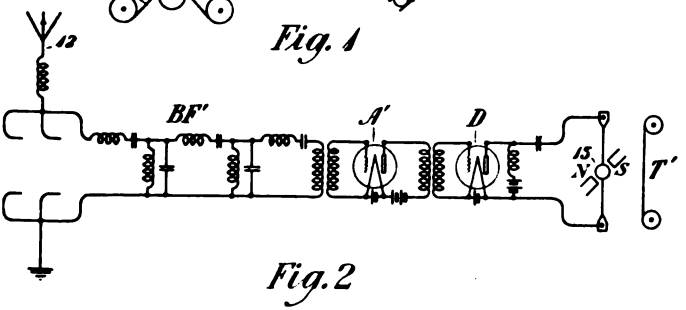

Fig. 2

NUMBER 1,494,347—High Speed Radio System

HIGH SPEED RADIO SYSTEM for accommodating a large volume of traffic over wave bands in close proximity to each other. The system consists on modulating a carrier frequency in accordance with signaling current produced at a speed sufficiently high so that the band of frequencies resulting from modulation will be too wide for transmission thru a sharply tuned selective circuit. The modulated band is selected with substantially negligible and uniform attenuation while adjacent frequencies lying outside of the bands will be sharply discriminated against by the use of band filters. At the receiving station a light stream is modulated in accordance with detected current and then the modulations recorded upon a medium sensitive to light.

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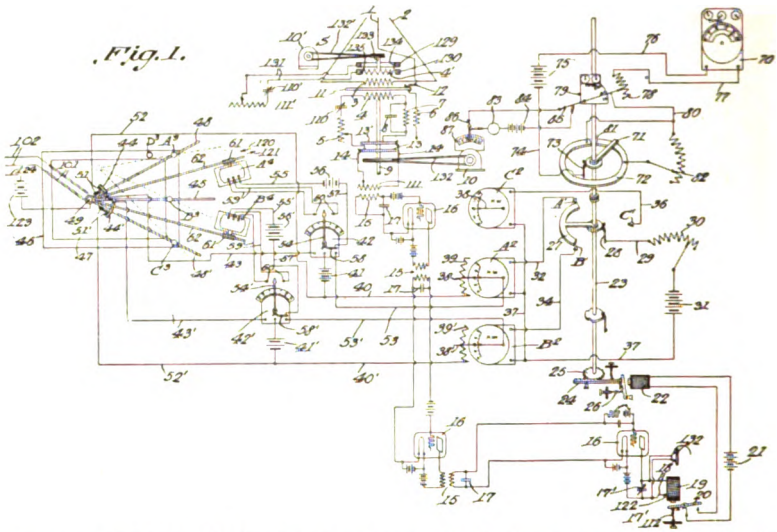
1,494,770—Walter W. Conners, filed June 12, 1919, issued May 20, 1924.



NUMBER 1,494,770—Method and Apparatus for Indicating the Geographical Location or Movement of Bodies

METHOD AND APPARATUS FOR INDICATING THE GEOGRAPHICAL LOCATION OR MOVEMENT OF BODIES by indicating in similitude or in intelligible signals the location of a moving body. The moving body carries a radio receiving station for the reception of signals from fixed wave-transmitting stations. The receiving station on the moving body is equipped with an indicating member which is controlled by received signals for indicating in similitude the signals received from the fixed stations. In this manner the course of the moving body is automatically plotted in accordance with the movement of the body with relation to the fixed stations.

1,494,803—Michael I. Pupin, filed September 17, 1915, issued May 20, 1924. Assigned to Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania.

ELECTRICAL TUNING apparatus by which the electrical resistance reaction which a conductor opposes to a simple harmonic electromotive force is rendered selective, that is to say, the conductor opposes a resistance reaction as small as desirable to an electromotive force of a given frequency, while at the same time this reaction may be made as large as desirable to an electro-

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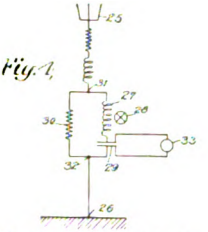
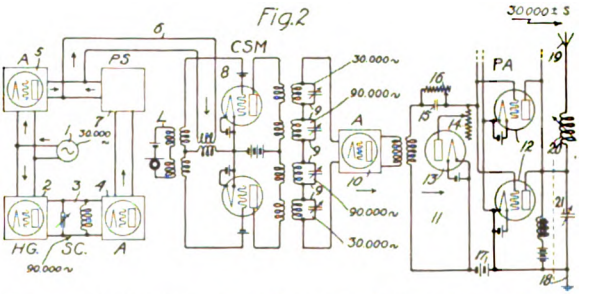
motive force of a different frequency or to electrical impulses. The invention is called a resistance compensator. It is shown as applied to a radio receiver for increasing the selectivity of the receiver by utilizing an antenna system of a resistance sufficient to dissipate the greater part of the incoming signaling energy. The incoming signaling energy of desired frequency is utilized to excite a source of energy by electromagnetic reaction and to transfer to the receiving antenna system a negative resistance reaction of high value as compared with the original resistance reaction of the conductor and of a frequency corresponding to the selected radio frequency to be received.



NUMBER 1,494,803—
Electrical Tuning

1,494,908—Raymond A. Heising, filed February 23, 1923, issued May 20, 1924. Assigned to Western Electric Company, Incorporated, New York, N. Y.



NUMBER 1,494,908—Amplification of Modulated Waves

AMPLIFICATION OF MODULATED WAVES for securing economy in operation of high power systems. The amplification system may be applied to a system which includes a source of carrier waves, a device for deriving from the source a triple frequency component, amplifying this component and combining it with waves from the original source to produce a peaked wave; a

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carrier suppression modulating device for modulating the peaked wave thus produced in accordance with a low frequency wave; a voltage step-up or low power amplifier for the modulated wave; and a vacuum tube power amplifier having a grid so negatively polarized that the amplifier is actuated only by the peak portion of each cycle of the modulated peaked wave. The repeater derives from the modulated wave a triple frequency component of large energy where the energy is chiefly in side frequencies based upon the fundamental frequency all of which energy is efficiently increased in amplitude thru the amplification system of this invention.

1,494,935—Edward O. Scriven, filed July 16, 1918, issued May 20, 1924. Assigned to Western Electric Company, New York, N. Y.

ALTERNATING CURRENT SOURCE where the frequency of oscillations generated is governed by a tuned oscillation circuit whereby the oscillator will generate substantially pure sine waves. An electron tube circuit is provided with the input and output circuit associated for feeding back energy from the output circuit to the input circuit. An additional circuit is interposed in the electron tube circuits for substantially preventing the feeding back of energy oscillations of any frequency other than that which is desired to generate by means of the electron tube oscillator.

1,495,537—Stephen F. Stafford, filed August 21, 1923, issued May 27, 1924. Assigned to Stafford Radio Company, Incorporated, of Massachusetts.

DOUBLE HELIX CAGE ANTENNA for indoor use in radio reception. The antenna consists of two helices wound in opposing relation with respect to each other with one nested within the other providing a strong cage antenna.

1,495,470—John F. Farrington, filed February 1, 1922, issued May 27, 1924. Assigned to Western Electric Company, Incorporated, of New York, N. Y.

HIGH FREQUENCY TRANSMISSION for multiplex and duplex operation where mechanical switching for two-way conversation is unnecessary. The system utilizes partial carrier wave suppression wherein the incoming and outgoing waves combine to produce an auxiliary carrier wave at each station. The transmitting system operates to transmit only feeble energy of the

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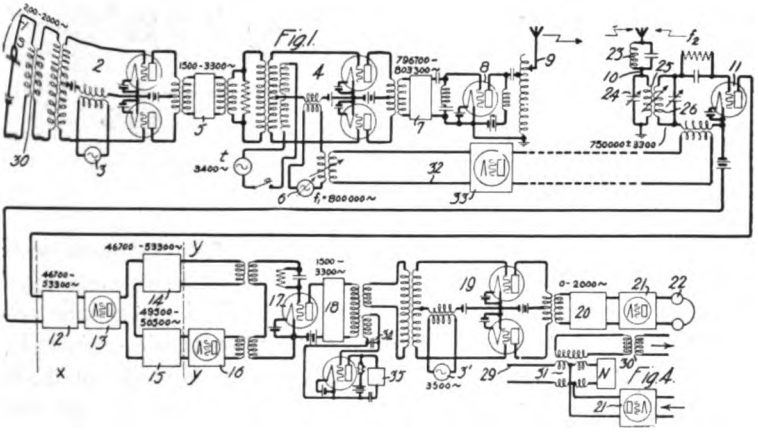
unmodulated component. The unmodulated component is separated from its accompanying side frequencies at a point in the system where the total energy is small relative to the transmitted energy. The energy of the selected unmodulated component is increased at the receiver and combined at the detector with the accompanying side frequencies to operate the receiving telephone.



NUMBER 1 495,470—High Frequency Transmission

1,495,511—Robert J. Fitzgerald, filed May 12, 1920, issued May 27, 1924. Assigned one-half to J. Arthur Fischer, New York, N. Y.

CONDENSER of the rotary variable construction, wherein the movable and stationary metallic plates have an outside insulating coating thereover. The insulating coating of each of the plates is adapted to contact one with the other preventing the matallic plates from shunting each other.

1,495,577—Francis W. Dane, filed April 3, 1920, issued May 27, 1924. Assigned to Wireless Specialty Apparatus Company, Incorporated, Boston, Massachusetts.

ELECTRICAL CONDENSER of the stacked type where alternate sheets of mica and metal foil are placed within a casing with a resilient spring member compressing the stack under a pressure of the order of thousands of pounds while the stack is submerged in insulating material which is solid at ordinary temperatures, but fluid at a temperature of the order of about 140 degrees centigrade. The construction described avoids the presence of air bubbles and forces the mica and foil in intimate contact with each other, securing high capacity but with high insulation properties.

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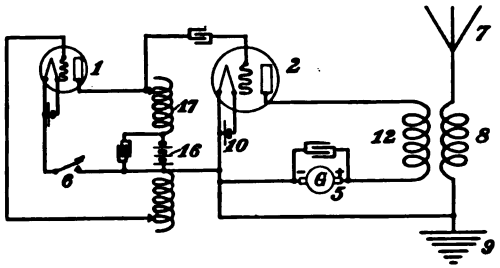
1,495,593—Lewis M. Hull, filed October 16, 1920, issued May 27, 1924.



NUMBER 1,495,593—Radio Telegraph Transmitting System

RADIO TELEGRAPH TRANSMITTING SYSTEM having an electron tube amplifier whose plate-cathode circuit is operatively coupled to the transmission circuit. A self-excited electron tube generating system is so connected to the grid-cathode circuit of the amplifier that the inherent intra-electrode capacity of the said grid-cathode circuit furnishes capacity reactance for the operation of the self-excited electron tube generating system. By this system the unavoidable intra-electrode capacity of the amplification system is transformed from a wasteful absorber of electrical power into an essential tuning element, and its use as a tuning element allows a reduction in the minimum capacitive reactance attainable in the oscillatory circuits, thus increasing the maximum frequency to which these circuits are resonant.

1,495,616—Frederick G. Simpson, filed May 23, 1923, issued May 27, 1924.

ART OF MEASURING THE VELOCITY OF TERRESTRIAL BODIES RELATIVE TO EACH OTHER by applying the Doppler-Fitzeau principle, force or effect, as manifested thru the medium of electromagnetic waves. A method is described for measuring the velocity of a terrestrial body with respect to another terrestrial body which consists in comparing the frequency of generation of a generator of electromagnetic waves, at rest with respect to one of the bodies, with the frequency of the electromagnetic waves radiated by such generator as measured at a point at rest with respect to the other body.

1,496,155—F. J. Fransson, filed November 8, 1919, issued June 3, 1924. Assigned to American Gas Accumulator Company, Elizabeth, New Jersey.

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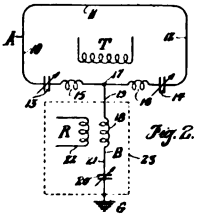
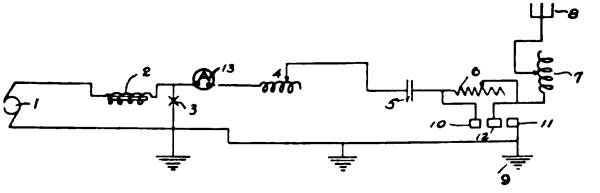


NUMBER 1,496,155—Art of Radio Communication

ART OF RADIO COMMUNICATION, in which a combination loop antenna and open antenna is employed for duplex operation. A transmitter is coupled to one of the antennas while the receiver is coupled to the other antenna with condensers arranged for balancing out interference between the transmitter and receiver during duplex operation.

1,496,214—L. F. Fuller and H. F. Elliott, filed May 22, 1919, issued June 3, 1924. Assigned to Mesne Assignments to Federal Telegraph Company, San Francisco, California.



NUMBER 1,496,214—Radiotelegraphy Transmission System

RADIOTELEGRAPHY TRANSMISSION SYSTEM, employing an arc with a signaling circuit for transmitting with a single wave. A circuit is employed to replace the antenna system during periods of no radiation. The circuit consists entirely of inductive and capacitive reactance neutralizing each other or tuned to the frequency of the system, and connected in series with the antenna. Thus, the total effect of this part of the circuit on the antenna circuit is negligible, and across the terminals of the inductance and capacity a minimum potential difference is obtained for currents of the correct frequency. By proper arrangement of a resistance or other consumer of energy and a switching means, the substitute circuit can then be made to have substantially the same characteristics as the antenna circuit, so that the load on the source remains the same, independent of choice of paths.

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In this manner the telegraph signals are produced without radiation of energy between signals.

1,496,243—F. S. McCullough, filed July 7, 1919, issued June 3, 1924. Assigned to G. L. Martin of Cleveland, Ohio.

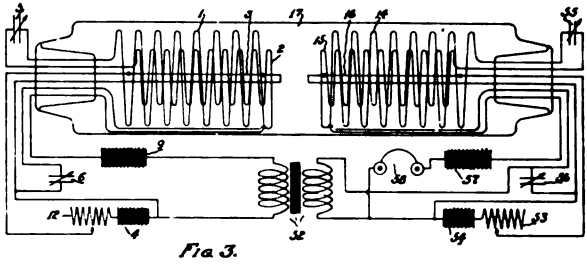

Fig 3.

NUMBER 1,496,243—Radio Telegraphy

RADIO TELEGRAPHY in which a compact form of apparatus is employed for the reception of radio signals particularly desirable for use on aircraft or in places where space is an extremely important factor. The collector coils which form the means for intercepting the radio frequency energy are wound within the evacuated tube which contains the electronic emission elements connected in the receiving circuit.

1,496,275—P. Laut, filed October 18, 1920, issued June 3, 1924.

RADIO TRANSMITTING APPARATUS employing an arc generator and a signaling circuit for producing uni-wave signals. The field coils of the arc are excited from an independent source. The antenna circuit is completely broken between signals enabling the arc to operate upon an auxiliary shunt circuit eliminating the compensation wave.

1,496,276—H. P. Rees, filed August 9, 1921, issued June 3, 1924.

RADIO TELEGRAPHY AND TELEPHONE RECEIVING SET containing constructural features for the reduction of size of the apparatus. The antenna tuning coil comprises a plurality of layers of wire, each layer consisting of a plurality of substantially plane turns. The axial length of the coils is small in relation to its mean diameter. The anode circuit of the set is reactively coupled with the grid circuit of the receiver.

1,496,311—J. H. Hammond, filed October 17, 1917, issued June 3, 1924.

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Receiving System for Radiant Energy, particularly adapted for radio control operation where positive operation of the control circuit is assured without the smothering or saturating effect which may be produced by powerful signals from a hostile station endeavoring to prevent weaker signals from the distant control station from having any appreciable effect upon the receiving system. To insure the positive operation of the control a plurality of electron tube detectors connected in parallel are arranged at the receiving circuit with the control connected in the common output circuit of each of the detectors.

1,496,371—C. B. Graves, filed August 23, 1922, issued June 3, 1924.

Condenser of variable capacity where the plates are formed in parallel relationship with dielectric sheets interposed between the conductive plates with means for exerting pressure against the plates for varying the capacity of the condenser.

1,496,671—H. Gernsback, filed February 24, 1923, issued June 3, 1924.

Detector of the crystal type where a sensitive crystal is enclosed by a casing which includes a feeler member in contact with the crystal. The feeler member comprises a central plate having plane opposite faces and side plates having a plurality of contact feeler points thereon which contact with a plurality of sensitive points on the crystal. By this construction a crystal detector may be placed in permanent adjustment.

1,496,745—H. S. Scott, filed June 13, 1923, issued June 3, 1924.

Potentiometer in which the variation in resistance takes place in accordance with a logarithmic law. An insulating band which carried the resistance element is relatively wide at a central section but decreases in width toward each end thereof on opposite sides of the central section. The resistance wire is wound upon this band whereby relatively large increments of adjustment are secured within the central section while relatively small increments of adjustment are obtained within the tapered portions on each side of the central section.

1,496,768—W. T. Booth and Walter A. Boyd, filed March 27, 1918, issued June 10, 1924. Assigned to Western Electric Company.

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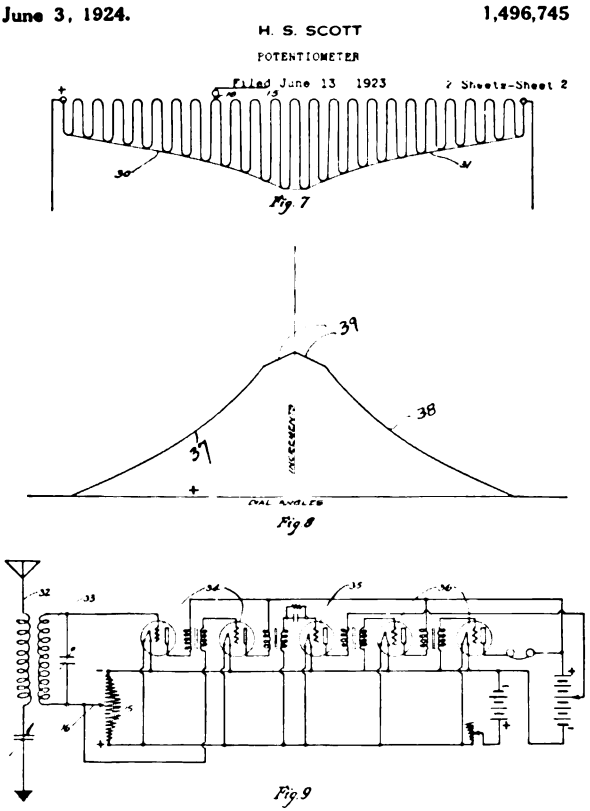
Vibration Reducing Mounting Device for Signaling Systems, where the electron tubes are supported on a horizontal panel, each end of which is mounted upon sponge rubber. The sponge rubber is carried by an intermediate supporting member and the intermediate support in turn carried by a second resilient mounting so that vibrations are prevented from reaching the tube support thru two separate damping means.



Number 1,496,745—Potentiometer

1,497,095—W. Dubilier, filed May 23, 1922, issued June 10, 1924. Assigned to Dubilier Condenser and Radio Corporation of Delaware.

Electrical Condenser of the "Micadon" variety where the alternate conductive and dielectric sheets are secured together under compression by means of a clip engaging each of the opposite ends of the condenser with a rivet passing thru the clips

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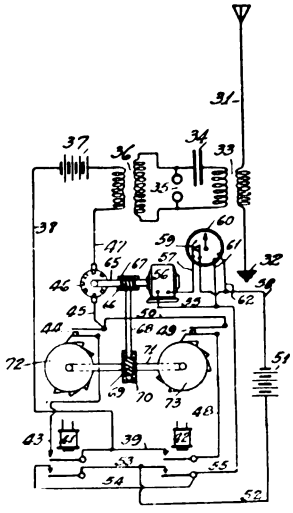
to secure the clips to the condenser and cause the clips to press tightly against the plates.

1,497,194—J. Nordern and A. Fetting, filed April 13, 1922, issued June 10, 1924.



NUMBER 1,497,194—Radio Alarm System

RADIO ALARM SYSTEM where a plurality of radio transmitting systems may be located in different houses to be automatically placed in operation by the happening of a condition precedent. The invention contemplates the protection of homes against burglary or fire by the installation of a radio transmitting station in the home. Upon the fusing of a fusable link in case of fire a radio transmitter is set into operation and controlled by means of a particular set of code wheels which will advise a central station that the particular home is in danger. In the case of a burglary a circuit might be opened by an intruder which places the radio transmitter in operation to advise the central police station. Each house will be identified by a particular code combination set out on the code wheels of the transmitter.

1,497,384—H. G. Saal, filed May 11, 1922. Issued June 10, 1924

RADIANT ENERGY RECEIVER, consisting of a crystal detector construction where the cat whisker is supported by a spherical member seated upon a spherically curved standard. A coil

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spring interior of the standard is secured to the surface of the spherical member enabling the member to be freely adjusted to different angular positions.

1,497,415—Phillips Thomas, filed April 21, 1921, issued June 10, 1924. Assigned to Westinghouse Electric and Manufacturing Company.

MOLDED OIL-FILLED CONDENSER, wherein alternate sheets of foil and solid dielectric are supported in a molded casing in position where a liquid dielectric may be filled in over the plates for increasing the insulating properties.

1,497,449—W. H. Kempton, filed October 16, 1917, issued June 10, 1924. Assigned to Westinghouse Electric and Manufacturing Company.

MOLDED CONDENSER comprising a series of stacked plates of alternate sheets of conducting material and an insulating material. The stack is wrapped with a compressible covering and the entire unit forced into a molded cover and held in compression.

1,497,927—A. Meissner, filed September 3, 1921. Assigned to Gesellschaft fur Drahtlose Telegraphie m.b.H., Hallesches, of Berlin, Germany. Issued June 17, 1924.

SENDING ARRANGEMENT, where the harmonics are suppressed by means of a grounded circuit arrangement in parallel with the grounded portion of the antenna. The grounded circuit is tuned to prevent the radiation from the antenna of harmonics of the fundamental frequency.

1,497,948—I. Shoenberg, filed April 9, 1918. Assigned to Radio Corporation of America, of Delaware. Issued June 17, 1924.

THERMIONIC DEVICE for securing an intermittent discharge of electrons from the filament. Alternating current is employed for heating the filament of a thermionic device. The temperature of the filament, provided its diameter is properly chosen, will then be proportional to the square of the current. Due to the thermal inertia, that is insufficient radiation from the filament, the temperature of the filament will lag behind the current. This property is utilized in various ways such as the reception of continuous wave radio signals. A choke coil saturated by a direct current is employed in series with the alternating current source for controlling the heating current to the filament.

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1,498,129—A. E. Spicer, filed June 5, 1923. Assigned to Thomas J. F. Coady, of New York. Issued June 17, 1924.

RADIO TELEPHONE RECEIVING APPARATUS, constructed in compact form and including a miniature pick-up loop. The tuning unit includes a primary coil having a tap substantially central of the length of the coil. The ends of the coil are connected by a shunt circuit through condensers. A rotary coil is positioned adjacent each end of the primary coil. One of the rotary coils is connected in series with the primary coil and the other rotary coil is connected between the central tap and a ground connection. The receiving apparatus is connected across the tuning system with suitable amplification for bringing in signals.

LIST OF RADIO TRADE MARKS PUBLISHED BY PATENT OFFICE PRIOR TO REGISTRATION

(The numbers given are serial numbers of pending applications)

184,263—"ALL-AMERICAN" for radio apparatus, in ornamental design. Rauland Mfg. Co., Chicago, Illinois. Claims use since October 15, 1919. Published May 6, 1924.

190,373—"CROWN" for radio apparatus, in ornamental design Crown Radio Mfg., Inc., New York, N. Y. Claims use since about November 1, 1922. Published May 6, 1924.

190,773—"SELECTOFORMER" for radio apparatus, Electrical Research Laboratories, Chicago, Illinois. Claims use since December 3, 1923. Published May 6, 1924.

191,386—"HETEROFORMER" tuning coil, Elvin Radio Co., Philadelphia, Pennsylvania. Claims use since on or about December 15, 1924.

191,387—"HETEROPLEX" for radio receiving sets. Elven Radio Co., Philadelphia, Pennsylvania. Claims use since on or about December 15, 1923. Published May 6, 1924.

191,423—"LISTEN-TO-THE-WORLD" Radio for radio receiving sets. B. & C. Radio Company, Inc., Boston, Massachusetts. Claims use since about December 1, 1923. Published May 6, 1924.

192,287—"SELECTOMETER" for wave traps and tuning instruments. Vesco Radio Shop, Oakland, California. Claims use since January 1, 1923. Published May 6, 1924.

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192,420—"SERENADA" for loud speakers. Racon Electrical Co., Inc., New York, N. Y. Claims use since December 21, 1923. Published May 6, 1924.

192,742—"AEOLUS" for radio receiving sets. H. M. Tower, Inc., West Haven, Connecticut. Claims use since November 1, 1923. Published May 6, 1924.

193,248—"RADIO" for radio apparatus, in ornamental design. Radio Sales Studio, Inc., Washington, D. C. Claims use since June 1, 1923. Published May 6, 1924.

183,901—"MIDGET" for radio sets. Midget Radio Company, Philadelphia, Pennsylvania. Claims use since May 4, 1922. Published May 6, 1924.

185,353—"FOUR-PHONE-PLUG" for radio phone plugs. The Barkelew Electric Mfg. Co., Middletown, Ohio. Claims use since September, 1922. Published June 10, 1924. Not subject to opposition.

170,132—"FROST RADIO" for radio receiving apparatus. Herbert H. Frost, Chicago, Illinois. Claims use since September 30, 1922. Published June 17, 1924.

170,133—"LIKE POSTAGE STAMPS, USED EVERYWHERE" for radio receiving apparatus. Herbert H. Frost, Chicago, Illinois. Claims use since August, 1921. Published June 17, 1924.

189,924—"RADIO UNITED" for radio receiving sets, in ornamental design. United Cigar Stores of America, New York, N. Y. Claims use since November 7, 1923. Published June 17, 1924.

191,499—"THE MOCKING BIRD" for radio receiving sets. Morrison Laboratories, Inc. Detroit, Michigan. Claims use since December 1, 1923. Published June 17, 1924.

191,584—"—RAMO—" for radio and wireless merchandise. Ramo Radium Products Co., New York, N. Y. Claims use since September 8, 1923. Published June 17, 1924.

193,844—"STRAND" for radio receiving apparatus, in ornamental design. Manufacturers Phonograph Co., Inc., New York N. Y. Claims use since January, 1921. Published June 17, 1924.

194,908—"RADIOSCOPE" for variable condensers. Chelten Electric Co., Philadelphia, Pennsylvania. Claims use since November 27, 1924. Published June 17, 1924.

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194,936—"NATIONAL" for radio transformers, in ornamental design. National Transformer Mfg. Co., Chicago, Illinois. Claims use since September 15, 1921. Published June 17, 1924.

195,143—"PHONODYNE" for radio receiving sets, in ornamental design. George A. Turner, Stockton, California. Claims use since March 1, 1924. Published June 17, 1924.

195,302—"WHAT'S IN THE AIR," a monthly magazine published by the Reincke-Ellis Co., Chicago, Illinois. Claims use since August 1, 1923. Published June 17, 1924.

195,317—"THURMAN" for vario couplers. Charles L. Thurman, Philadelphia, Pennsylvania. Claims use since February, 1924. Published June 17, 1924.

195,830—"RADIO PROGRESS," a magazine. Oxford Press, Providence, Rhode Island. Claims use since March 15, 1924. Published June 17, 1924.

193,794—"APCO" for battery chargers. Apco Mfg. Co., Providence, Rhode Island. Claims used since 1910. Published May 27, 1924.

193,846—"MOON" C-2A, in ornamental design, for radio receiving sets. Moon Radio, Inc., Long Island, New York. Claims use since May, 1922. Published May 27, 1924.

193,908—"EIRACO" for radio receiving sets, in ornamental design. Elgin Radio, Inc., Elgin, Illinois. Claims use since October 17, 1922. Published May 27, 1924.

194,119—"NICKELETTE" for radio receiving sets. Novelty Radio Mfg. Co., St. Louis, Missouri. Claims use since February 23, 1924. Published May 27, 1924.

183,318—"NIGHTINGALE" for radio receiving sets. The Cleveland Apparatus Co., Grafton, Ohio. Clains use since May 1, 1923. Published May 27, 1924.

187,490—"SHACTON RADIO PRODUCTS" in ornamental design, for radio apparatus. Harold M. Schwab, Inc., New York, N. Y. Claims use since August 6, 1923. Published May 27, 1924.

189,384—"MELCO" for radio receiving sets and parts thereof. Amsco Products, Inc., New York, N. Y. Claims use since July 12, 1922. Published May 27, 1924.

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189,473—"OPERADIO" for radio receiving sets. The Operadio, Inc., Chicago, Illinois. Claims use since April 15, 1922. Published May 27, 1924.

191,711—"MACKLINE" in ornamental design, for radio antennas and tube-projecting apparatus. Franklin H. Mackenzie, doing business as Mack Co., Philadelphia, Pennsylvania Claims use since July 16, 1923. Published May 27, 1924.

193,080—"SUR-RAD" for radio receiving sets and apparatus therefor. Hugh Stevenson, Rochester, New York. Claims use since October 7, 1924. Published May 27, 1924.

194,630—"LAFAYETTE" for radio apparatus. R. E. Thompson Mfg. Co., Jersey City, New Jersey. Claims use since February 7, 1924. Published May 27, 1924.

194,713—"AIR SCOUT" in ornamental design for radio receiving sets. Kilbourne and Clark Mfg. Co., Seattle, Washington. Claims use since about December 1, 1923. Published May 27, 1924.

195,254—"DERESNADYNE" for radio receiving sets and parts. Andrews Radio Co., Chicago, Illinois. Claims use since March 1, 1924. Published June 17, 1924.

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COUPLED CIRCUITS," PUBLISHED IN THE "PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS," JUNE, 1924

By

E. Leon Chaffee

All letters L, C, R, and γ occurring in Part I should have the subscript unity.

In some equations the subscript is outside the exponent as θ^2_1. These typographical errors do not change the meaning, and the above symbol should read in the usual form as θ_1^2.

Page

300 In equation (1-1) there should be no minus sign before C in the second bracket of the denominator.

304 Equation (4-1), R^2 should read R_2.

305 In equation (5-a) the fraction in the first parenthesis of the denominator should be $\frac{2}{\theta_1^2}$.

The third line below equation (5-a) should read

$$1-\frac{\theta^2}{\theta_1^2}=\beta_1 \text{ and } 1-\frac{\theta^2}{\theta_2^2}=\beta_2$$

306 In equation (7-1) R_2 should be R_1^2.

Equation (7-a) should read

$$\beta_2=\frac{\tau^2\beta_1}{\gamma_1^2\theta^2+\beta_1^2}$$

Italics beside equations (7-1), (7-a), (7-b), should read *Adjustment of secondary reactance to give max. I_2. Equation of Max. 1-2 line.*

307 In equation (8-a) M_1 should be γ_1.

In the second line below (8-a) parenthesis should be after I_2.

309 In equation (14-a) I^2 should be I_2.

312 Omit parenthesis in third line above Figure 6.

313 In the second line below equation (17-a) γ_1 should be $\frac{\gamma_1}{\gamma_2}$.

In the third line below equation (17-a) $\tau=\frac{\gamma_1}{\gamma_2}$ should be $\tau=1$.

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314 In the fifth line below Figure 7 $y = y_o$ should read $\lambda = \lambda_o$.

In the fifth line from the bottom of the page change y to λ.

315 In the fourth, fifth, and seventh lines above Figure 8 change y to λ. Insert "and" before "moving" in the fifth line above Figure 8.

317 On this page every y should be changed to λ except in the eleventh line from the bottom where $y = 1$.

320 In the second line below equation (20) "effects" should be "affects."

321 In equation (25-a) and in the line below and also in the third line of the second paragraph k should be K.

Equation (29) should read $\beta_2 = 1 - \dfrac{1}{\left(\dfrac{\bar{\lambda}_2}{\bar{\lambda}}\right)^2} = 1 - \dfrac{\theta^2}{\bar{\theta}_2^2}$

322 Equation (33) should read $\bar{\gamma}_{l2} = \dfrac{\bar{R}_2}{\bar{L}_2\,\omega_o} = \gamma_{l2}\dfrac{L_2}{\bar{L}_2} - H\dfrac{L_p}{\bar{L}_2}\gamma_{lp}$

In the fifth line from the bottom of the page the last R_2 should be $\bar{R}_2$, and in the second line from the bottom R_2 should be $\bar{R}_2$.

323 Case 1 should read "Special Case of," etc.

In equation (36-a) $\gamma_{l2}\theta$ should be $\bar{\gamma}_{l2}\theta$.

The $\overline{X}_2$ in the third and fourth lines from the bottom, in equation (37-1), and in the notation to the right of (37-1) should not be in Claredon Type.

The italicized notation to the right of equations (37-1) and (37-a) should read

"$\overline{X}_2$ to give max. I_2. Equation of max. 1-2 line.
$\overline{\beta}_2$ to give max. I_2. Equation of max. 1-2 line."

324 The parenthesis in the denominator should be closed at the end of equation (38-1), also the parenthesis should be closed after 6 on the last line.

329 Equation (47) should read

$$[\gamma_1\bar{\gamma}_2(\gamma_1+\bar{\gamma}_2)\bar{\theta}_2^2+\bar{\gamma}_2+\bar{\tau}^2\gamma_1]\theta_1^4+[\bar{\gamma}_2^2(\gamma_1+\bar{\gamma}_2)\bar{\theta}_2^2 - 2\bar{\gamma}_2(1-\bar{\tau}^2)]\theta_1^2\bar{\theta}_2^2+\left[\frac{\bar{\gamma}_2^2}{\gamma_1}(\gamma_1+\bar{\tau}^2\bar{\gamma}_2)\right]\bar{\theta}_2^4=0$$

332 Equation (48) should read $\theta^2 = \dfrac{\gamma_1+\bar{\gamma}_2}{\dfrac{\gamma_1}{\bar{\theta}_2^2}+\dfrac{\bar{\gamma}_2}{\theta_1^2}}$

333 In equation (49-a) in the first set of brackets $\bar{\tau}^1$ should be $\bar{\tau}^2$.

$\dfrac{1}{\theta^2}$ is the coefficient before the second set of brackets.

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334 Insert θ before $L_1 \omega_o$ in the second parenthesis of the denominator of equation (50-a).

In both forms of equation (50-a) E in the numerator should be E_1. In the last line of the second form of equation (50-a) the coefficient before the last set of brackets is $\frac{1}{\theta^2}$.

In the second and fourth lines below (50-a) $\frac{\lambda_2}{\lambda_o}$ should be $\frac{\lambda_2}{\lambda_o}$.

335 The second line should read "system at point P_1," etc.

338 In equations (45-2) and (46-2) $\bar{R}_{21}$ and $\bar{X}_{21}$ should read R_{21} and X_{21}.

339 In the first line S should be s.

$\bar{R}_{21}$ and $\bar{X}_{21}$ in equations (55-1) and (56-1) should be R_{21} and X_{21}. In the second line of equation (58), γ should be γ_1.

340 In equation (60) the first brace under the second square root sign of the denominator should read

$$\left\{\gamma_2 - \frac{s\gamma_p}{\theta_2^2} + \frac{\tau^2\gamma_1}{\gamma_1^2\theta^2 + \left(1 - \frac{\theta^2}{\theta_1^2}\right)^2}\right\}^2$$

341 In the first line above Figure 22 and the last line on the page, the number 23 should be inserted between the numbers 22 and 24.

342 In the second line from the top of the page in place of "below" insert "by equations (59) and (60)."

346 In the eleventh line from the bottom of the page insert "in" after "drop."

349 The italics after equation (61) should read "*Intersection of θ line. Tip of point.*

The fifteenth line from the bottom of the page should read "so that the settings for the tip of the point are," etc.

350 In the third line above Figure 31 $\bar{\gamma}_1 = -\gamma$ should be $\gamma_1 = -\bar{\gamma}_2$. In the first line above the same figure insert a semicolon after the first inequality. The second inequality should read $|\bar{\gamma}_3| < \gamma_1$.

In the first line above equation (63) substitute $\bar{\gamma}_2$.

354 In the fifth line from the top of the page k should be K. In the seventh line fraction should read $\frac{1-\theta^2}{\theta}$.

The second line of the second paragraph should contain

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the fraction $\frac{I_2 L_2}{I L_1}$ in place of the one given.

In the seventh line from the bottom of the page change γ_2 to $\gamma_1'' = .05$.

357
358 In the "Nomeclature" all k's should be K's.
359

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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 12 | OCTOBER, 1924 | Number 5

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A METHOD OF MEASURING RADIO FIELD INTENSITIES AND ATMOSPHERIC DISTURBANCES*

By

L. W. Austin

(Physicist)

and

E. B. Judson

(Associate Radio Engineer)

(Bureau of Standards, Washingtgn, D.C.)

(Communication from the International Union of Scientific Radio Telegraphy)

The Telephone Comparator

The following method of measuring telephone currents has been developed to take the place of the older telephone shunt method.

The apparatus is shown in Figures 1 and 2. Here 1 represents the 1,000-cycle tuning fork generator, manufactured by the General Radio Company, Cambridge, Massachusetts, Type 213. This generator operates on 6 volts and delivers a current of about 15 milliamperes thru 100 ohms. This has been found considerably superior to any audio-frequency electron tube generator which has been constructed, on account of relative freedom from harmonics, and ease in suppressing residual effects. To prevent hearing the fork directly, the oscillator is generally placed outside the receiving room.

Figure 2 represents the circuits making up the telephone comparator. The voltage divider has a resistance of 100 ohms, with two dials having steps of one and ten ohms. The milliammeter used is of the thermo-element type, having a range of 0 to 25 milliamperes, the current used being generally 10 milliamperes, giving a fall of potential across the voltage divider of one volt. The current to the telephones passes thru resistances R_1 and R_2, which are large enough to make variations in the telephone impedance

*Received by the Editor, August 8, 1924.

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nearly negligible. Generally 200,000 ohms are used. In this way it is easy to calculate the current in the telephones without accurate knowledge of their impedance. The telephone terminals are connected to a double-pole, double-throw switch which permits them to be connected either to the comparator, or to the radio receiving circuit.

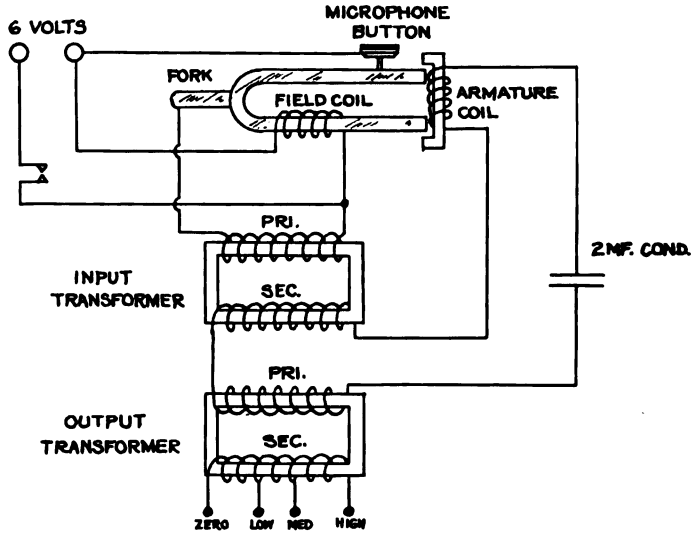

FIGURE 1—General Radio Company Type 213 1000 ∞ Audio Oscillator

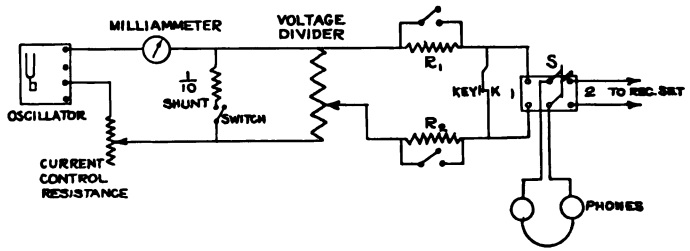

FIGURE 2—Telephone Current Comparator

MEASUREMENT OF SIGNALS

The process of measuring signals in the case of beat reception is as follows: the telephones are connected to the receiving set, and the radio frequency circuits are tuned to give beat reception, and the heterodyne or autodyne condenser adjusted to give the same beat note as the audio oscillator. If the signals are at all strong, the telephones are shifted forward on the head until the

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received signals are just audible since the comparison of sounds is much more accurate if they are weak. The telephones are then thrown to the telephone comparator side and the voltage divider adjusted until the telephone sound is of the same intensity in the two positions of the switch, the comparator currents being broken up to simulate signals by means of the key K. In this way the telephone currents can be measured with an accuracy of from 3 to 5 percent.

For the measurement of spark signals and atmospheric disturbances, no matching of the note of the comparator is possible, which, of course, materially reduces the accuracy of the measurements.

For the measurement of very weak signals it is desirable to reduce the current thru the voltage divider without correspondingly reducing the current in the milliammeter. This may be done by placing a one-tenth shunt across the voltage divider at S.

Electric Field Intensity

In order to make absolute measurements of the field from the transmitting stations, it is necessary to know (1) the resistance of the receiving antenna at the given wave length, (2) the radiation height (effective height) of the receiving antenna, and (3) to calibrate the telephone current as measured on the telephone comparator in terms of antenna current.

(1) The resistance (R) of the antenna is best obtained by the resistance-variation method, using a tube oscillator as the source of power and measuring the current in the antenna with a thermo-milliammeter or thermo-element and galvanometer (see Bureau of Standards, Circular Number 74).

(2) The radiation height (h_r) of the receiving antenna can be measured by measuring the current received from a station one to ten wave lengths away, of known transmitting current (I_s), radiation height (h_s) and distance (d), using thermo-element and galvanometer, from the formula, expressed in wave length λ and also in frequency f,

$$I_r = 120\pi \frac{I_s h_r h_s}{\lambda d R} = 40 \times 10^{-5} \pi \frac{I_s h_r h_s f}{d R}$$

in meters, ohms, kilocycles, amperes,

$$h_r = \frac{I_r \lambda d R}{120 \pi I_s h_s} = \frac{I_r d R}{40 \times 10^{-5} \pi I_s h_s f}$$

If this measurement cannot be made, the approximate height h_r may be taken as ½ the average height of the flat top (equivalent

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to the proposal by Dellinger.[1] If a coil antenna (loop) is used instead of an elevated antenna, its effective height can be calculated, disregarding the effect due to capacity to earth (antenna effect) from the formula.

$$h_r = 2\pi \frac{area \times number\ af\ turns}{\lambda} = 6.67 \times 10^{-6} f \cdot area \times turns$$

(3) The telephone current as measured in the telephone comparator can be calibrated in terms of antenna current from a signal of known intensity or better from a local oscillator as described later. The constancy of this calibration over long periods is very satisfactory, provided the receiving circuits are tuned as follows: Both antenna and secondary circuits are first tuned with very loose coupling so that there is no reaction between them. The coupling is then closed until the best signal is obtained, returning the secondary as may be required, but leaving the *antenna tuning unchanged.* This method, while not giving the loudest possible signals, can be depended on to give constant and perfectly reproducible results. The detuning of the secondary at least at the longer wave lengths in autodyne reception prevents the introduction of appreciable coupling resistance in the antenna.

Experiment shows that the telephone current, I_t, in beat reception is proportional to the square root of the watts, W, in the receiving antenna, and if E be the total emf. on the antenna, h the radiation height, e the electric field intensity, and R the resistance

$$W = \frac{E^2}{R} = \frac{h^2 e^2}{R} \quad \text{and}$$

$$e = \frac{\sqrt{WR}}{h}$$

since $\sqrt{W} = B\ I_t$ where B is the sensitivity constant of the apparatus.

The field intensity $e = \dfrac{B\, I_t \sqrt{R}}{h}$

where B is to be determined from observations on signals of known strength or from the local oscillator.

If it is desired to construct the apparatus for the measurement of signals, the following may be of service:

A suitable two-dial voltage divider can be made from any

[1] "Principles of Radio Transmission and Reception with Antenna and Coil Aerials," J. H. Dellinger, Bureau of Standards Scientific Paper, Number 354; 1919.

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dial resistance box by reconnecting the units and tens as shown in Figure 3 and cutting out the other dials or keeping them at zero.

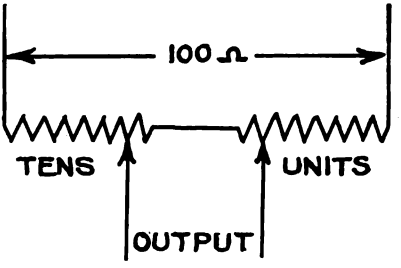

FIGURE 3—Connection for Voltage Divider

The thermo-milliammeter in the comparator may be replaced with some considerable reduction in cost by a Western Electric Company's vacuum thermo-element and General Radio Company's 10-ohm pointer galvanometer. A 35-ohm thermo-element shunted by six or seven ohms is convenient as this can be used without shunt in the local oscillator to be described later. The thermo-element galvanometer combination is calibrated by direct current which must be reversed thru the thermo-element heater and the mean deflection taken if a difference in reading is shown on reversal. Another possible inexpensive combination for measuring the comparator current is a Weston thermo-galvanometer (115 milliamperes full scale) and proper current transformer, tho the deflection will be small.

Two Lavite units of 100,000 ohms each are convenient for the resistances in the telephone leads.

A simple and compact tube oscillator can be constructed, using a low-current tube (UV-199) with dry cell "A" battery and "B" battery and the circuit shown in Figure 4. It is important that the whole oscillator be enclosed in a copper-lined box with tight-fitting cover. The copper box is grounded and also the lower end of the 1-ohm shunt, preferably a straight piece of Number 36 "Advance" wire. It is also important that the coil L_4 (Figure 5), to which the secondary of the receiving set is coupled, should be next the ground, for even a small amount of capacity coupling will produce large errors in the measurements. If a closed coil antenna (loop) is used instead of an elevated antenna the 1-ohm shunt from the oscillator is best placed in the middle of the loop. The thermo-element used in our laboratory is a Western Electric Company's type of 35 ohms heater resistance and the galvanometer is a General Radio Company's 10-ohm pointer instru-

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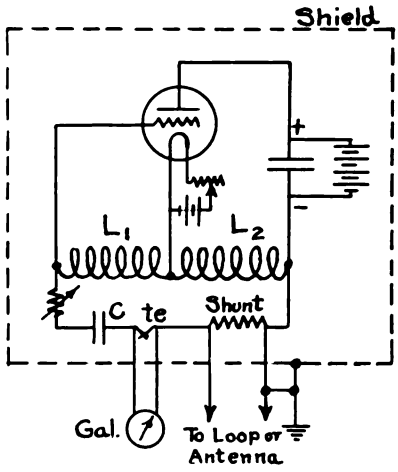

FIGURE 4—Shielded Oscillator

ment. The galvanometer can in most cases be kept outside the shielded box. As calibration on one wave length in the general range of observation is all that is required, if the oscillator is not to be used for other purposes, the condenser C for the sake of compactness can be of the fixed type. For work with trans-Atlantic stations, $L_1 = 25$ mh., $L_2 = 5$ mh., and $c = 0.002$ μf. This will give a wave length of about 15,000 m. The radio frequency current in the thermo-element should be between one-half and three milliamperes. This may be regulated by the coupling between the coils and by the filament current, or by resistance.

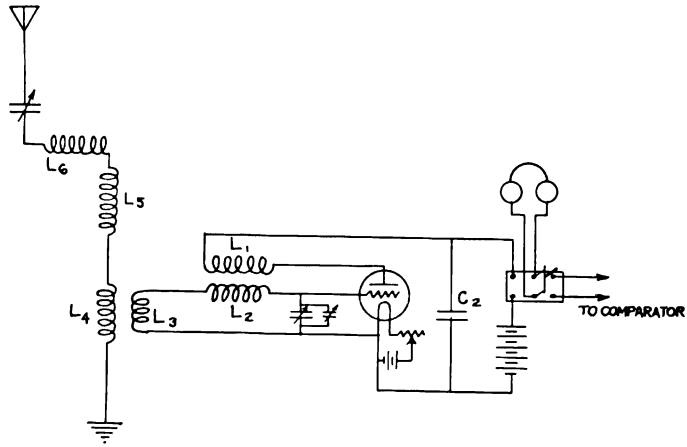

FIGURE 5

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Example of Calibration

Suppose a current of one milliampere is sent thru the 1-ohm shunt of the tube oscillator. This impresses 1 millivolt on the antenna, so if the antenna resistance including coils is 75 ohms, the power in the antenna

$$W = \frac{E^2}{R} = \frac{1\times10^{-6}}{75} = 1.33\times10^{-8} \text{ watts}$$

$$\sqrt{W} = 1.15\times10^{-4}$$

Experiments in the laboratory show that the strength of the signal current in the telephones with this connection, for a certain potentiometer setting, is, for our apparatus, approximately five times the strength of the current with the comparator used normally with 200,000 ohms in series with the telephones.

Now if at the same time the telephone comparator with its voltage divider set at 13 ohms gives the same intensity of sound in the telephones as the tube oscillator with the secondary properly coupled to the antenna, there will be a voltage on the telephone circuit from the comparator of 0.13 volts. If the telephones have an impedance of approximately 30,000 ohms and the added resistance in the telephone circuit is 200,000 ohms, the telephone current

$$I_t = \frac{1.30\times10^{-1}}{2.3\times10^{5}} = 5.65\times10^{-7} \text{ amp.}$$

Since the telephone current is proportional to the square root of the antenna watts in beat reception

$$\sqrt{W} = B\,I_t$$

where B is the sensitivity constant of the receiving set and

$$B = \frac{\sqrt{W}}{I_t} = \frac{1.15\times10^{-4}}{5.65\times10^{-7}} = 204$$

The Receiving Set

The autodyne receiving set used in the measurements is shown in Figure 5. The inductances L_1, L_2, L_4, L_5, and L_6 are coils of 30 mh., inductance and L_3 is 5 mh. The 30 mh. coils are wound with litz wire (48 strands, Number 38) on flat wooden spools. The wire space is 21 mm. long with an inside diameter of 72 mm. and outside diameter of 150 mm. The 5 mh. coils are similar but with an outside diameter of winding of about 100 mm. These have a resistance at the trans-Atlantic wave lengths of a little less than one ohm per millihenry. While it is desirable to have the coils in the antenna of low resistance, this is of no con-

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sequence in the secondary circuit, and solid wire can be used if desired, since resistance can be compensated by increasing the back coupling. Flat type coils are convenient for long-wave work but are not suitable below 2,000 meters and may be replaced thruout the set by honeycomb or other good coils which may be used at all wave lengths. The coil L_3 is attached to flexible leads for convenience in varying the coupling to the antenna coil L_4, which should be nearest the ground. The secondary variable condenser has a maximum capacity of 0.003 μf, in parallel with a 0.0005 μf. condenser used for exact tuning. The variable antenna condenser has a capacity of 0.005 μf., while the capacity of the fixed stopping condenser C_2 is 0.002 μf. To guard against variations in tube sensitivity with use, two tubes of the same sensitivity are chosen and connected so that either one can be thrown into the set. In this way a check can be kept on the constancy of the regular observing tube by occasional comparison with the spare.

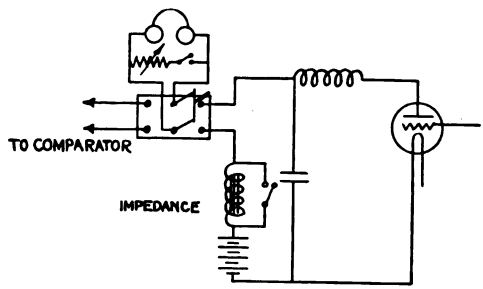

FIGURE 6—Circuit for Measuring Strong Signals

SPECIAL METHODS OF USING THE TELEPHONE COMPARATOR

1. MEASUREMENT OF VERY STRONG SIGNALS

Some difficulty has been experienced in measuring the strength of strong American stations with the telephone comparator, because it is generally necessary to keep the telephones at a considerable distance from the ears, in order to obtain a perfect comparison. This condition often allows outside noises to distract the observer's attention.

The following arrangement permits keeping the telephones on the ears, as in ordinary reception, even in the case of very strong signals.

An impedance equivalent to that of the telephones is connected in series with the telephones in the plate circuit of the receiving

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set with a switch provided to short circuit it when necessary. The telephones are shunted with a variable non-inductive resistance which can be switched out of the circuit when not in use. With the equivalent impedance in the circuit the shunt across the telephones is varied until the received signal becomes as weak as desired. The shunted telephones may then be switched alternately from the receiving set to the comparator and measurements made in the usual manner with the telephones directly on the ears.

Comparatively weak signals can be measured thru heavy atmospheric disturbances with this method, provided the disturbances are not too continuous, as it eliminates the deafening effect of the strong disturbances.

For ordinary signal measurements the impedance coil is shorted out by a single pole single throw switch and the shunt disconnected from the telephones. Figure 6 shows the schematic diagram.

2. Measurement of Atmospheric Disturbances

The early method of determining the strength of atmospherics was unsatisfactory. In this the disturbances were compared with the 1,000-cycle generator output in the same manner that radio signals are compared. The difference in the sound of the two sources and the variations in the strength of the disturbances made measurements difficult and inaccurate. Often, disturbances were too strong to measure.

The method now in use applies the voltage from the output of the comparator directly to the plate circuit of the receiving set, the high resistances being cut out. Atmospheric disturbances and the 1,000-cycle signal from the comparator can then be heard simultaneously, and the strength of the comparator signals is adjusted until when sending at about twenty words per minute they become just readable thru the disturbances. Hand sending or an automatic transmitter placed in the tuning fork oscillator line can be used to produce the signals. Experiments in the laboratory show that the strength of the signal current in the telephones with this connection, for a certain voltabe divider setting, is for our apparatus as nearly as can be measured five times the strength of the current with the comparator used normally with the 200,000-ohm resistance in series with the telephones. Therefore, the field intensity corresponding to the voltage divider setsing necessary to read thru the atmospheric disturbances, when multiplied by five, is taken as the field strength of the disturbances. Figure 7 shows the connections used.

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3. Measurement of Amplified Signals

It is nearly impossible to measure the strength of signals with the telephone comparator in the usual way when audio-frequency amplification is used. When the 1,000-cycle tuning fork is running, the amplifier generally amplifies the residual current induced in the receiver from the leads of the comparator, sufficiently to drown a radio signal of moderate strength. It has been found necessary therefore to open the "zero" side of the tuning fork oscillator leads when the comparison switch is thrown to the radio signal and close it when thrown to the comparator. This can be done by means of an ordinary telegraph relay placed close to the tuning fork oscillator and controlled by a switch which can be thrown simultaneously with the comparison switch. The signals can then be measured in the usual manner.

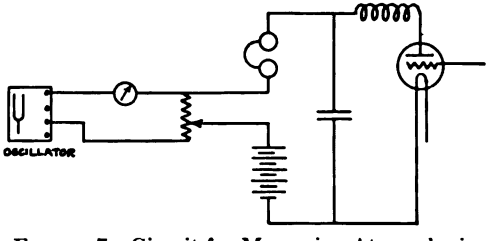

Figure 7—Circuit for Measuring Atmospherics

4. Apparatus for Measuring the Radiation Height of Transmitting Stations

The apparatus described was intended for measurements on stations having antenna currents of 40 amperes to 250 amperes, masts more than 60 m. high, and wave lengths of 6,000 m. and above. The receiving loop is 24.4 m. long, 9.15 m. high, and wound with 15 turns of Number 16 (diameter = 1.29 mm.) hard drawn copper wire, the turns being spaced approximately 15 cm. apart. The frame is swung between two portable masts 14 m. high and 30 m. apart at a distance of about one wave length from the transmitting station to be measured, preferably with level ground between, and no trees in the immediate vicinity of the loop. The direct current resistance of such a loop is approximarely 12 ohms.

The indicating apparatus consists of a Leeds and Northrup portable galvanometer having a coil resistance of approximately 20 ohms, and requiring about 20 microvolts for a deflection of one division; and a Western Electric Company's vacuum thermo-

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element, Type 5a, having a heater resistance of about 35 ohms and a thermocouple resistance of about 10 ohms. The thermo-element can be calibrated by direct current provided the current is reversed and the mean of the two galvanometer readings taken, if any difference is shown. A typical calibration shows 10 mm. galvanometer deflection for 0.71 milliamperes in the heater of the thermo-element, and 140 mm. for 2.58 milliamperes.

The transmitting station usually makes one-minute dashes, with intervals of one minute between. This enables the resistance of the loop to be measured by inserting known additional resistance and noting the reduction in deflection. A station with 180 m. masts transmitting with 100 amperes at 10,000 m. will give 1.5 milliamperes in the loop, or 48 mm. galvanometer deflection assuming a 63-ohm total loop resistance; and indicates a radiation height of the transmitting station of 120 m. A station with 90 m. masts, transmitting at 6,000 m. with 50 amperes will give about 30 mm. on the galvanometer scale.

The calculations expressed both in wave length and frequency are made from the formula

For Two Antenna:

$$I_r = 120\pi \frac{I_s h_s h_r}{\lambda d R} = \frac{40\times 10^{-5}\pi I_s h_s h_r f}{d R}$$

$$\text{then} \quad h_s = \frac{I_r \lambda d R}{120\pi h_r I_s} = \frac{I_r d R}{40\times 10^{-5}\pi h_r I_s f}$$

For Loop Receiving:

If the radiation height of the receiving loop is

$$h_r = 2\pi \frac{area \times turns}{\lambda} = 2\pi \frac{NHL}{\lambda} = 6.67\times 10^{-6}\pi NHL f$$

then the radiation height of the transmitting antenna will be

$$h_s = \frac{I_r \lambda^2 d R}{240\pi^2 I_s NHL} = \frac{I_r d R}{26.7\times 10^{-10}\pi^2 I_s NHL f'}$$

when I_r is received current in loop,

I_s the transmitting current,

h_s the transmitting radiation height.

N, H, L are respectively the number of turns, height, and length of the loop,

λ the wave length,

f the frequency,

d the distance from transmitting station, and

R the total resistance in the loop circuit.

All quantities are expressed in meters, kilocycles, ohms, and amperes.

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SUMMARY: The method of measuring radio signals here described depends on the determination of the equality of the sound intensities in the telephones from the signal and from a known alternating current preferably giving the same tone. The measured local current is taken from a thousand-cycle tuning fork oscillator and the electromotive force impressed on the telephones is controlled by a voltage divider. The errors of measurement may be under good conditions less than five percent, tho in case of spark signals and atmospheric disturbances when the two tones cannot be matched it will be very much greater. Special arrangements are described for measuring very strong signals and atmospheric disturbances.

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DISCOVERY OF A PLACE WHERE NO STATIC COULD BE HEARD IN AUGUST*

By

R. H. Marriott

(Navy Yard, Bremerton, Washington)

In August, 1921, we found a place at Pop Point on Thorne Arm in Southeastern Alaska, sixteen miles (25.6 km.) east-northeast of Ketchikan, where no static was heard for six days, altho numerous radio stations on the Pacific Ocean and on the Pacific Coast could be heard complaining to each other and asking for repetition of messages because of interference from static.

Tests made with directive antenna, at Astoria, Oregon, for several days in July, 1920, and at odd times later, indicated that the chief static centers affecting receiving at Astoria were probably Mt. Jefferson, Mt. Wilson, Mt. Hood, Mt. St. Helens, Mt. Adams, and Mt. Rainier; Mt. Rainer apparently being the chief center, altho farther from Astoria than Mts. St. Helens and Adams.

Tests made at Bremerton, Washington, two or more times per day from July 26, 1920, to September, 1921, gave Mt. Rainier as the chief center about 90 percent of the time. Mt. Rainier is about fifty miles (80 km.) from Bremerton. Some of the tests were made simultaneously at Astoria and Bremerton, giving cross bearings on Mt. Rainier. In addition to the static maxima received from the direction of Mt. Rainier, occasional storm static centers became apparent first in the southwest and traveled over Bremerton, producing an all-direction effect when near by, followed by a dying-out directional effect in the northeast. From November 15th to April 15th the static was usually too indefinite to judge direction.

Mt. Rainier (another name for which is Mt. Tacoma) rises abruptly from a nearly sea level elevation. To the west of the mountain is a comparatively low land area over which cool air from the Pacific can move toward the mountain and be heated by the sun's rays and warm earth and obtain some dust and considerable moisture. This more or less saturated and dust-bear

* Received by the Editor, April 10, 1924

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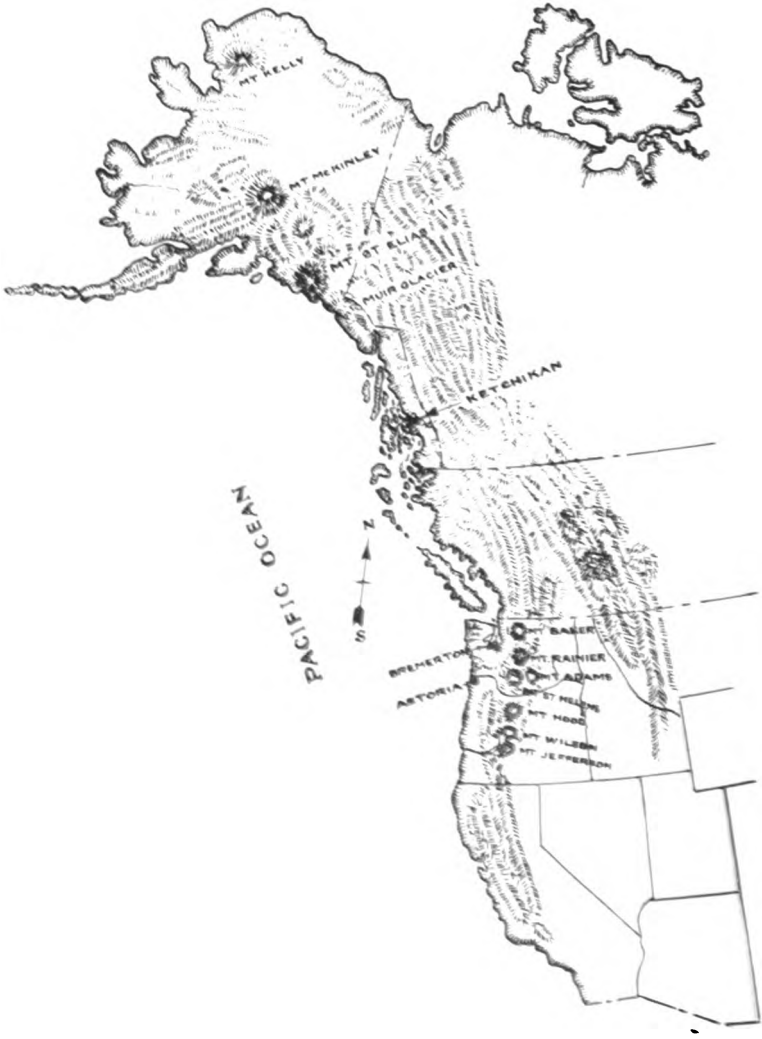
ing air may then come quite abruptly in contact with Mt. Rainier, which is 14,471 feet (4,410 m.) high and capped by perpetual glaciers. If the temperature of the air west of Rainier reached as high a value as air reaches in the eastern part of the United States, Mt. Rainier might be the most violent static center in the United States. However, our air averages probably 15° F. (8° C.) cooler than eastern air, and we are bothered less by static. Probably it should also be mentioned that Mt. Rainier is classed as an extinct volcano and that visitors to the top occasionally report that some gases are still given off.



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The other peaks previously mentioned are high and cold on top but are not so high as Rainier or do not bear the same relation to warming-up areas for air. Numerous other mountains in the territory to be considered are not snow-covered in summer or else the warm, moisture-bearing air probably does not come in contact with them, and they probably screen off the effect of static and radio signals instead of producing static. The static-producing peaks also apparently have a screening effect for radio signals and probably as well for static disturbances coming from points beyond them.

Passing northward along the Pacific Coast at sea, the effect of static apparently decreases. Passing northward from Tacoma, thru Puget Sound, which is separated from the Pacific by the Olympic peninsula, and then north back of the islands along the coast of Canada and the coast of southeastern Alaska, the static effect apparently decreases on the average until a point is reached in Icy Straits southeast of Mt. St. Elias, where the ocean-going vessels leave the inside passages and go out to sea across the Gulf of Alaska towards Cordova. From that point, just inside of Icy Straits, the static interference apparently gets stronger. Directional measurements from Cordova indicated maxima in the direction of Mt. St. Elias. Mt. St. Elias rises abruptly from the sea to 18,024 feet (5,500 m.) This mountain and the vicinity are covered by many glaciers. However, air from the Gulf of Alaska is probably never very warm because the Pacific Ocean is comparatively cool. Also there have been reports of quite violent static by radio operators located at Yakutat, which is near Mt. St. Elias.

Returning from Cordova we selected a location sixteen miles (25.6 km.) east-northeast of Ketchikan.

In choosing this site we gave particular attention to the fact that the study of natural static is commonly interfered with by man-made static. Not only is it sometimes difficult to distinguish man's static from nature's static, but man-made static sometimes drowns out the natural static. For example: our readings on Mt. Rainier static were somewhat confused by a "Cottrell" precipitator of dust from the smoke of the Tacoma, Washington smelter. That device uses many kilowatts in producing a multitude of sparks. Mt. Rainier's cold top, east of Tacoma, apparently serves to condense warm moisture and dust-bearing air and to precipitate static discharges, thus making little and big sparks. The man-made precipitator west of Tacoma charges the plates of air condensers to high voltages by rectifying high

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voltage alternating current thru synchronous rotary spark gap rectifiers. Each air condenser consists of an insulated negative wire in a grounded positive concentric vertical tube one foot (0.305 m.) in diameter. The smoke passes upward thru the tube in the static field between the wire and tube where the dust is precipitated to the wire and wall of the tube, the gases passing on and out the stack. The charges on the tube and wire do not confine themselves to remaining static electricity or even kinetic charges in the form of minute sparks to minute particles of dust, but they jump across five or six inches (12 to 15 cm.) quite frequently and irregularly, between the wire and tube. Also the high voltage alternating current occasionally flashes over. Altogether this is an extreme case of man-made static, some of which is quite like natural static, and all of which is capable of drowning out natural static, radio messages, and broadcasts over a considerable area. Tacoma radio listeners are between the beautiful natural precipitator (which they call Mt. Tacoma) and the economical man-made precipitator on Puget Sound.

The site we chose at Pop Point was sixteen miles (25.6 km.) from the nearest high tension line or other likely source of man-made electrical interference. Having arrived at that site, a directional receiving loop and antenna similar electrically to those we had been using for a year was set up and used. No static was heard for the six days from August 9th to 15th. Nothing could be heard but radio stations and the faint sounds produced by the tubes and batteries. We had avoided both man-made and nature-made static. Ships and shore stations as far north as Cordova, as far west as Honolulu and as far south as southern California were heard. We used the regenerative detector and two stages of audio frequency. However, we did not hear any radio stations east of us or south of us as far inland as Seattle or Mt. Rainier. We were apparently screened from that section. The testing party consisted of Mr. R. L. Barbur and I. Mr. Barbur, in addition to being able to make the direction tests, is an experienced radio operator. On the seventh day static was heard just as we were leaving and after the loop part had been dismantled. This static was apparently from a moving storm which blew in the following day from the southwest.

Leaving this point we went to Ketchikan, where we heard static on a small antenna. From there we went to Sitka and, mounting our directional apparatus on a forty-foot (12.2 m.) gas boat, we cruised around among the islands. The static in the vicinity of Sitka was weak and no definite directions were noted.

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The ships' radio operators traversing the inside passages in the island fringe along the coast of Canada and Southeastern Alaska have reported for years their inability to receive radio stations in certain spots. These reports have shown numerous combinations of ability to hear certain stations and not others, which sometimes reverse as the vessel proceeds. Broadcasting has brought out similar reports relative to the screening effect of inland mountains, and our receiving about Bremerton has always been handicapped by such mountains as the Olympics west of us. Tne screening effect of the high rugged land of this country is apparently much greater than the diminishing effect of the lower land on the east coast.

Those islands of Western Canada and Southeastern Alaska are numerous. The larger islands are mountain high. Some islands carry peaks over three thousand feet (910 m.) in height. The channels between the mountains as a rule are comparatively narrow. In some cases ocean-going vessels pass thru waterways that are not over four hundred feet (120 m.) wide. In many cases standing on the deck of a vessel and endeavoring to see the tops of the mountain, it is necessary to look up as you would look up from the street to see the tops of the buildings in an American city. Also the mainland of Alaska and Canada is a similarly rough and rugged arrangement of mountains and valleys. Tall trees are characteristic of the entire section.

My assumption is that the no-interference point sixteen miles (25.6 km.) east-northeast of Ketchikan was separated from permanent static centers by just enough rough and rugged topography to be screened from those static centers while that point was not similarly screened from the Pacific Ocean and Pacific Coast.

SUMMARY: The origin of natural static near mountain peaks is described and compared, in its mode of production and character, with that produced by smoke precipitators.

Some mountains are believed to be soucres of static, while others act as absorbing shields to signal and static. On these bases, a certain point near Ketchikan, Alaska, was found where static was not heard in midsummer, while signals were received from some directions but not from others.

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THE SUPER-HETERODYNE—ITS ORIGIN, DEVELOPMENT, AND SOME RECENT IMPROVEMENTS*

By

Edwin H. Armstrong

(Marcellus Hartley Research Laboratory, Columbia University New York)

The purpose of this paper is to describe the development of the super-heterodyne receiver from a wartime invention, primarily intended for the exceedingly important radio telegraphic direction-finding service in the Signal Corps of the American Expeditionary Force, into a type of household broadcasting receiver, which, with our present vision, appears likely to become standard.

The invention of the super-heterodyne dates back to the early part of 1918. The full technical details of this system were made public in the Fall of 1919.[1] Since that time it has been widely used in experimental work and is responsible for many of the recent accomplishments in long distance reception from broadcasting stations. While the superiority of its performance over all other forms of receivers was unquestioned, very many difficulties rendered it unsuitable for use by the general public and confined it to the hands of engineers and skilled amateurs. Years of concentrated effort from many different sources have produced improvements in vacuum tubes, in transformer construction, and in the circuits of the super-heterodyne itself, with the result that at the beginning of the present month there has been made available for the general public a super-heterodyne receiver which meets the requirements of household use.

It is a peculiar circumstance that this invention was a direct outgrowth to meet a very important problem confronting the American Expeditionary Force. This problem was the reception of extremely weak spark signals of frequencies varying from about 500,000 cycles to 3,000,000 cycles, with an absolute minimum of adjustments to enable rapid change of wave length. The

*Presented before The Institute of Radio Engineers, New York, March 5, 1924. Received by the Editor, April 26, 1924.

[1]Proceedings of The Institute of Radio Engineers, February, 1921. Presented December 3, 1919.

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technical difficulties of this problem are now so well known that it was not necessary to consider them. Round in England and Latour in France, by some of the most brilliant technical radio work of the war, succeeded in producing radio frequency amplifiers covering the band from 500,000 to 1,000,000 cycles and tho covering a much more limited band, amplifiers operating on 2,000,000 cycles had been constructed. These results had been accomplished by the use of vacuum tubes and transformers of a minimum capacity. As this apparatus was used in the highly important intelligence services, all information was carefully guarded. When the United States entered the war, the facts that it was necessary to produce sensitive receivers for short wave lengths and that tube capacity would prove the bar to a straightforward solution of the problem were not known in this country. As a result, no attention was paid to the capacity in the type of vacuum tube which was adopted, and while the tube met the requirements of the lower frequencies admirably, it was impossible to use it effectively for the frequencies of importance in the direction-finding service.

During the early part of 1918, thru the courtesy and energy of General Ferrié and his staff, the American Expeditionary Force was supplied with apparatus of French manufacture. It was quite apparent, however, that this source of supply could not be a permanent one, and a solution of the problem became essential. During the early part of 1917, I had made a careful study of the heterodyne phenomena and their effect on the efficiency of amplification. With this work freshly in mind, the idea occurred to me to solve the problem by selecting some frequency which could be handled by the tubes available, building an effective amplifier for that frequency, and then transforming the incoming high frequency to this readily amplifiable value by some converting means which had no low limit; preferably the heterodyne and rectification. The principles and advantage of this method were explained in a paper presented before this INSTITUTE[3] and are now so well known that no further explanation is required here.

After much experimental work, an eight-tube set was constructed consisting of a rectifier tube, a separate heterodyne oscillator, three intermediate frequency amplifiers, a second

[2] PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, April, 1917. Presented October 4, 1916.

[3] This amplification is based on the ratio of the voltage applied to the second detector to the voltage at the loop terminals. The intermediate frequency amplification is unknown.

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rectifier or detector, and two audio frequency stages. The intermediate frequency stages were coupled by tuned air-core transformers set for a frequency of about 100,000 cycles, with an adjustment for controlling the regeneration. The amplification of voltage measured at the input of the second detector with the amplifier just below the oscillating point, was about equivalent to a radio frequency amplification of 500.[4] This is illustrated in Figure 1 and the arrangement of its circuits in Figure 2. It gave satisfactory results except that the inclusion of a regenerative control on the intermediate frequency amplifier made skilled handling necessary, as the adjustment of the frequency of the oscillator changed the plate current of the detector tube and this, in turn, varied the resistance which that tube introduced into the amplifier system and upset the regenerative adjustment.

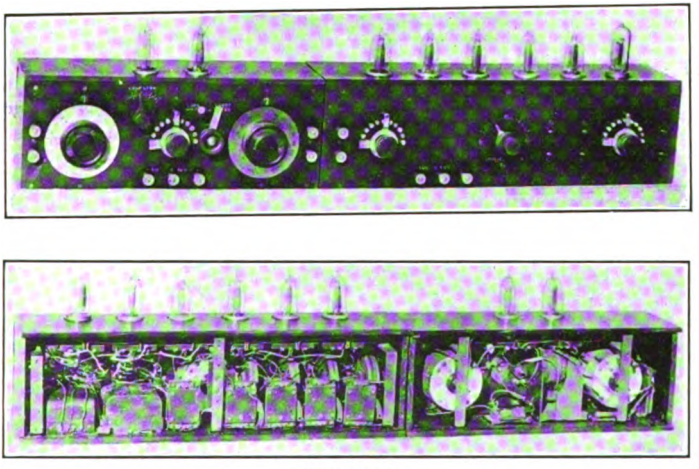

FIGURE 1

The Armistice ended development at this point, but in the fall of 1919, for the purpose of determining the results which could be obtained by pushing the super-heterodyne method of reception to the limit, a resistance-coupled intermediate frequency amplifier consisting of five high-μ (amplification factor) tubes was constructed. The voltage amplification of these five stages was probably between 5,000- and 10,000-fold. While greater amplification could have been obtained, the sensitiveness of a set composed of a two-tube frequency converter, a five-

[4] PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, February, 1921. Presented December 3, 1919.

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tube intermediate frequency amplifier, a detector, and one-stage of audio, was such that on a three-foot (one-meter) loop, the sole criterion of reception was simply whether the signal was stronger than the atmospheric disturbances.

The sensitiveness of the super-heterodyne was demonstrated during the winter of 1919-1920, when the spark signals from amateur stations on the West coast and telephone signals from destroyers in Southern waters were received in the vicinity of New York on a three-foot (one-meter) loop. Probably the most striking demonstration of the capabilities of the method occurred in December, 1920, when Paul F. Godley, at Ardrosson, Scotland, received the signals of a large number of amateur stations located in the United States, many of them being spark stations. The super-heterodyne used by Godley consisted of a regenerative tube for the rectifier, a separate oscillator, four stages of re-

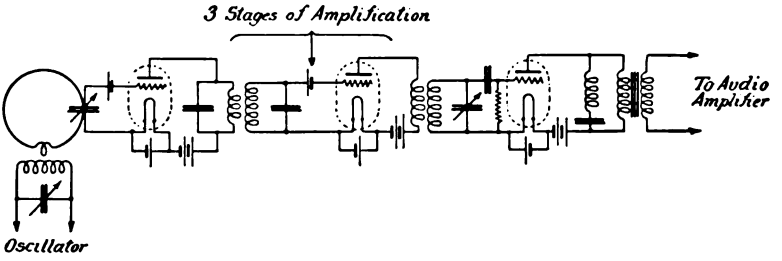

FIGURE 2

sistance-coupled intermediate frequency amplification, a second rectifier, and two stages of audio. While it is difficult to state definitely the actual voltage amplification obtained, it appears to have been between 3,000- and 5,000-fold.[5]

With the coming of the broadcasting art, and with the great increase in the number of stations and the consequent interference, the super-heterodyne began to take on a new importance —an importance which was based not on its superior sensitiveness nor on its selectivity, but on the great promise which the method offered in simplicity of operation. It was, and still is, the standard practice to furnish the public with receivers equipped with a variety of tuning adjustments for the purpose of amplifying the desired band of radio frequencies and excluding all others. As a matter of fact, many more adjustments than are on receivers should be used—more than could be placed in the hands of the average user. It would obviously be of the

[5] Based on standard previously described. This is without the second heterodyne which was used in receiving continuous waves.

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greatest importance if in some way these tuning adjustments could be made in the laboratory by skilled engineers and sealed, leaving some relatively simple adjustment for the hands of the operator. The super-heterodyne offered the ideal solution. This solution lay in the construction of an intermediate frequency amplifier which would amplify a given frequency and a band 5,000 cycles above and below it and which would cut off sharply on either side of this desired band. The adjustments necessary to accomplish this could all be made by skilled men, and the only operations left for the user would be the two adjustments necessary to change the incoming frequency down to the band of the amplifier—adjustments which are not dependent on each other, which are of extreme simplicity, and which can be made equally well by the novice or the engineer. To determine just what could be accomplished along these lines, the writer, working in conjunction with Mr. Harry Houck, constructed during the spring of 1922, a set designed for the maximum usable sensitiveness and selectivity. The set-up consisted of one radio frequency stage (non-tuned transformer) a rectifier tube, and oscillator tube (used as a separate heterodyne), a three-stage iron-core transformer-coupled intermediate frequency amplifier designed to cover a a band of 20,000 to 30,000 cycles, a second detector tube, and two-stage of audio frequency amplification. UV-201 tubes were used. The set without the audio frequency amplifier is illustrated in Figure 3 and Figure 4. To prevent the intermediate frequency amplifier from oscillating, each stage was shielded separately. The use of a radio frequency stage ahead of the first detector possesses a number of advantages, but the chief one is in eliminating the reaction between the loop circuit and the oscillator circuit. Experience with the original type had shown that when an oscillator of ordinary power was used, it was necessary to couple it rather closely with the loop circuit in order to insure a sufficiently strong heterodyne current. This close coupling affected the tuning of both circuits, an adjustment of one changing the setting of the other. To avoid this trouble and to produce a system wherein a station could always be tuned in on exactly the same settings, a single stage of radio frequency amplification (non-tuned transformer) was used, and the oscillator was coupled into this transformer. This arrangement eliminated the reaction, reduced the radiation to a minimum and, in addition, removed the damping of the first rectifier from the loop circuit and improved its selectivity.

The results obtained with this set were about as expected.

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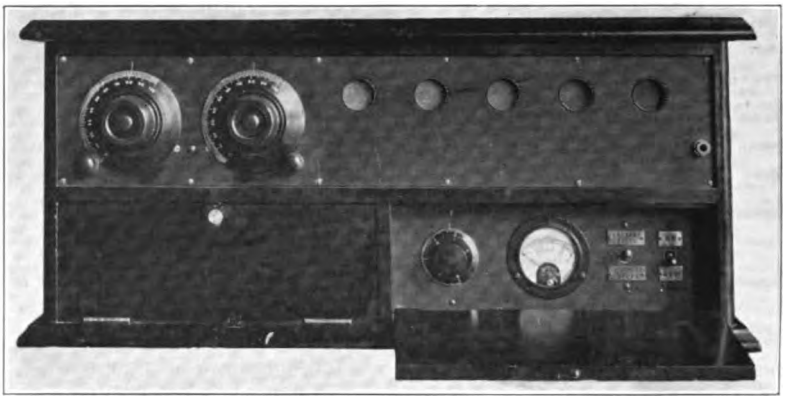

FIGURE 3

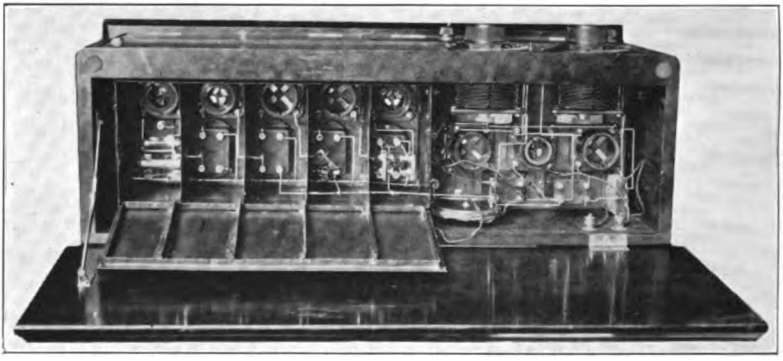

FIGURE 4

On a three-foot (one-meter) loop, the factor determining the reception of a station was solely whether the signal strength was above the level of the atmospherics. The selectivity was such that stations which had never been heard before on account of blanketing by local stations, were received without a trace of interference. While the performance of the set was much superior to any other receiver, it was apparant that the cost of construction and maintenance was prohibitive. The single item of a ten-ampere filament current will give some idea of the size of the storage battery and auxiliary apparatus required.

With the coming of the low filament consumption, or dry battery type of tube, the possibilities of producing a super-heterodyne for household use were tremendously improved. The set of Figure 3 was remodelled for the WD-11 tube, and its sensitiveness was brought to about the same value as obtained with the

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storage battery tubes. This was a long step forward, but still the cost was prohibitive.

It has been apparent ever since the question of the application of the super-heterodyne to broadcasting had been considered, that there were too many tubes performing a single function which were quite capable of performing a double one. The most outstanding case is that of the separate heterodyne oscillator. In view of our knowledge of the self-heterodyne, it appears quite obvious to perform the first rectification by means of a self-heterodyne oscillator and thereby save a tube. As a matter of fact, this was one of the very first things tried in France, but, except for very short wave lengths, it was never very successful when a high intermediate frequency was necessary. The reason was this: If a single tuned oscillating circuit was used, the mistuning to produce the proper beat caused a loss of signal strength which offset the gain of a tube. If two tuned circuits were used on the oscillator, one tuned to the signaling frequency and the other arranged to oscillate at the heterodyne frequency, then on account of the relatively small percentage difference in frequency a change in the tuning of one circuit changed the tuning of the other. The solution of this problem was made by Houck, who proposed an arrangement so simple and so effective that it completely solved the problem. Houck proposed to connect two tuned circuits to the oscillator, a simple circuit to the frequency of the incoming signal and a regenerative circuit adjusted to oscillate at such a frequency that the second harmonic of this frequency beating with the incoming frequency produced the desired intermediate frequency. The general arrangement is illustrated by Figure 5.

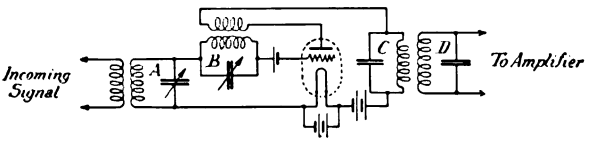

FIGURE 5

In the diagrammatic illustration, circuit A is tuned to the incoming signal, circuit B is tuned to one-half the incoming frequency plus or minus one-half the intermediate frequency, and the circuits C and D are both tuned to the intermediate frequency. The operation of the system is in line with ordinary self-heterodyne action. By reason of the asymmetrical action of the tube, there are created in the circuits a variety of har-

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monics. The second harmonic combines to produce beats with the incoming signals of the desired intermediate frequency. the tube rectifies them to produce the desired intermediate frequency and. thru *C* and *D*, the new frequency is supplied to the amplifier. On account of the fact that circuits *A* and *B* are tuned to frequencies differing by approximately 100 percent. a change in the tuning of one has no appreciable effect on the tuning of the other. This arrangement solved the oscillator problem and. in addition, practically eliminated radiation.

The next step in the reduction of the number of tubes was to make the radio frequency amplifier perform the function of amplifying intermediate frequency as well. This can be done with none of the difficulties inherent in audio frequency amplification, as the very small amplitudes of voltage handled by the first tube preclude the possibility of the grid becoming positive with respect to the filament. The general arrangement of circuits for carrying this out is illustrated by Figure 6. In this

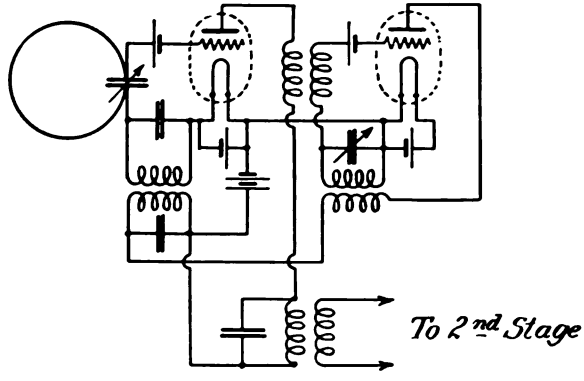

FIGURE 6

arrangement the signals received by the loop are amplified at radio frequency by the first tube and applied to the grid of a second harmonic oscillator by means of an untuned radio frequency transformer. The combined signaling and heterodyning currents are then rectified by the second tube, producing a current of the intermediate frequency which is applied to the grid of the first tube, amplified therein, and passed on to the second stage of the intermediate amplifier. A more practical method of carrying out this idea is illustrated in Figure 7. In this arrangement, a secondary of the first intermediate frequency transformer is connected to the grid of the first tube and in parallel

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with the loop circuit. Otherwise, the arrangements of Figures 6 and 7 are identical. The parallel type of circuit arrangement eliminates a variety of reactions which would give rise to oscillations of various frequencies and in addition, prevents the reception of long wave signals by the intermediate frequency amplifier. When this development had been completed, improvements in the design of the intermediate frequency transformers made it possible to obtain with two stages all the amplification which could be used.

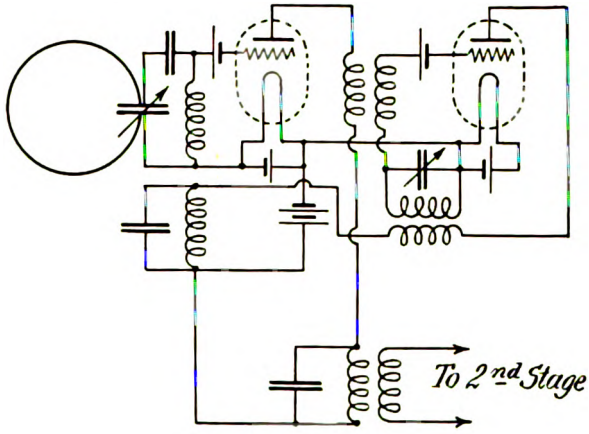

FIGURE 7

On account of the high amplification, signals from local stations overload the second rectifier and introduce distortion. Control of the amount of intermediate frequency amplification is essential. While there are numerous methods equally effective, the simplest one appears to be the control by means of the filament temperature of the second intermediate frequency amplifier.[6]

The features just described were all incorporated in the receiver, which is illustrated in Figures 8 and 9. The set measured 18 by 10 by 10 inches (45.6 by 25.4 by 25.4 cm.) and was completely self-contained—the batteries, loop antenna, and speaker mechanism being enclosed in the box. The results were highly satisfactory, and loud speaker signals (at night) in the vicinity of New York were obtained from stations in Chicago and Atlanta. It demonstrated that not only could a household receiver

[6]Although some form of potentiometer type of control of the voltage (a. c.) applied to the grid of one of the amplifier tubes would obviously be better, the simplicity of the filament control has many advantages in manufacture.

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of the super-heterodyne type be built, but that the first practical solution of the portable set was at hand.

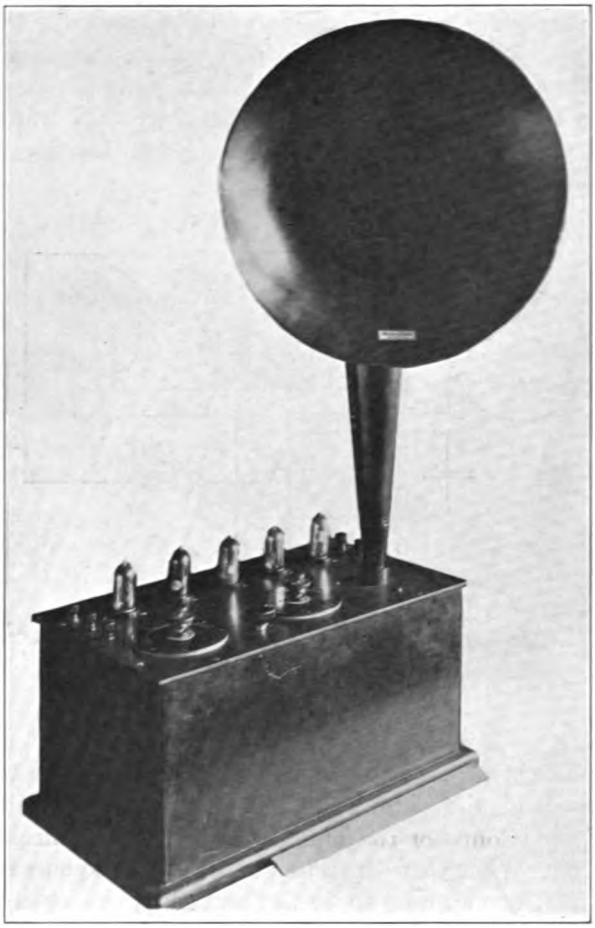

FIGURE 8

In this form, the capabilities of the set were brought to the attention of the Westinghouse Electric and Manufacturing Company and the Radio Corporation of America a little over a year ago. Its possibilities were instantly visualized by Mr. David Sarnoff, who immediately took steps to concentrate the resources of the research laboratories of the Radio Corporation of America, the Westinghouse Electric and Manufacturing Company, and the General Electric Company on this new development. From that point on it passed into a new phase—that of placing an in-

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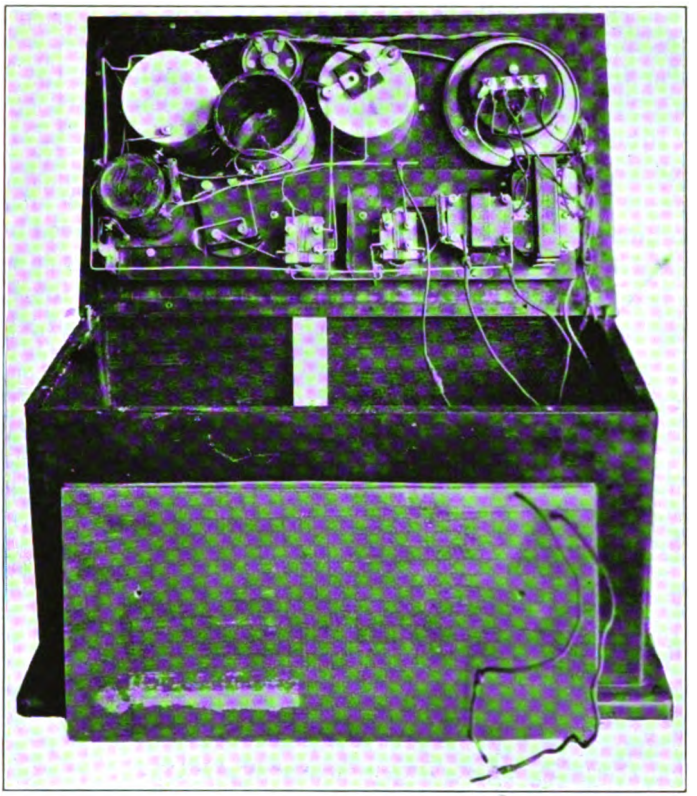

FIGURE 9

vention in a commercial form. In the limited time available, this was a most extraordinarily difficult proposition, and credit for its accomplishment is due to the untiring efforts on the part of the engineers of the above organizations. Many improvements and some radically new ideas of designs have been introduced, but it is the privilege of those responsible for them to present these. In the final development, an additional stage of audio frequency amplification was added in order to insure operation within steel buildings, particularly those within the city limits where signals are relatively very weak compared to suburban locations. This makes a six-tube set, but six tubes can be readily operated on dry batteries and the increase in sensitiveness is well worth the extra tube.

Some idea of the sensitiveness and the ease of operation of the set illustrated in Figures 10 and 11 may be gathered from an incident during the trans-Atlantic broadcasting tests of November and

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December, 1923. On December 1st, two ladies, neither having any technical radio knowledge, received loud speaker signals from station 2LO, London, England. This was accomplished at Merrimac, Massachusetts, with the set and loop illustrated in Figures 10 and 11 and probably constitutes a record for the first radiophone reception from Europe with a portable receiver. With the same set and a three-foot (one-meter) loop, loud speaker signals from broadcast stations on the Pacific Coast were received in the vicinity of New York on an average of three or four times a week. The factor determining reception was simply whether the signal strength was above the level of the atmospheric disturbances.

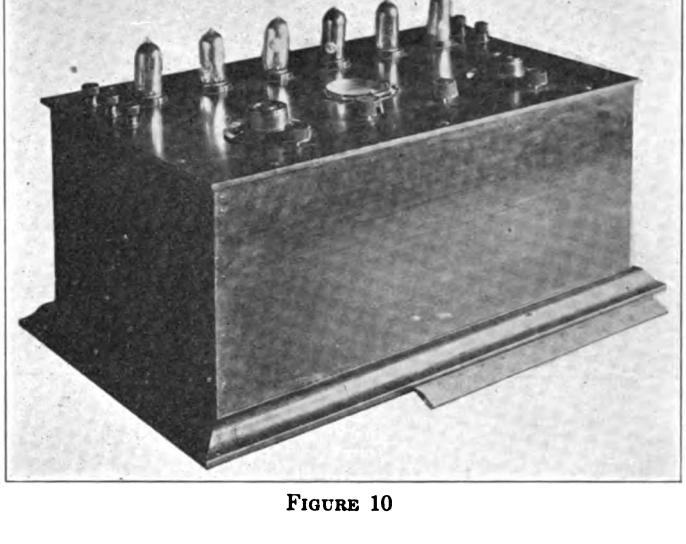

FIGURE 10

The type of super-heterodyne described is now available to the public in the two forms illustrated in Figures 12 and 13. Each of these sets incorporate the arrangements herein described. Their sensitiveness is such that, with a two-foot (61-cm.) loop and an unshielded location, the atmospheric disturbances are the criterion of reception. Here we reach a milestone in the development of broadcast receivers, for no increase in the distance of reception can now be obtained by increase in the sensitiveness of the receiver. Unless the power of transmitting stations is increased we are about at the limit of the distance which can be covered. Future improvement of this receiver will lie along the line of selectivity and simplifying the construction.

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SUMMARY: This paper describes the development of the super-heterodyne receiver from a wartime invention into a commercial form of broadcast receiver apparatus now available to the general public. The success of the development is due to the low filament consumption vacuum tube and to the reduction in the number of tubes required by self-heterodyning, reflexing, and improvement in transformer design.

Instances are cited of trans-Atlantic and trans-continental reception of broadcast stations by completely portable sets constructed in accordance with the methods described.

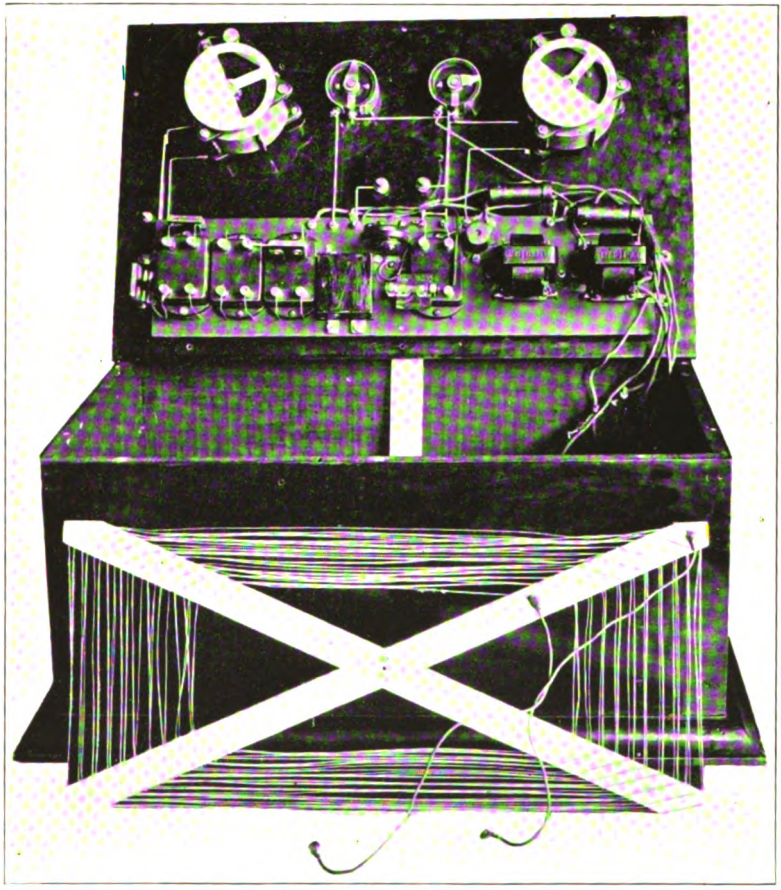

FIGURE 11

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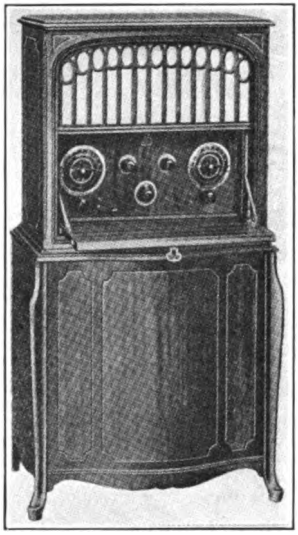

FIGURE 12

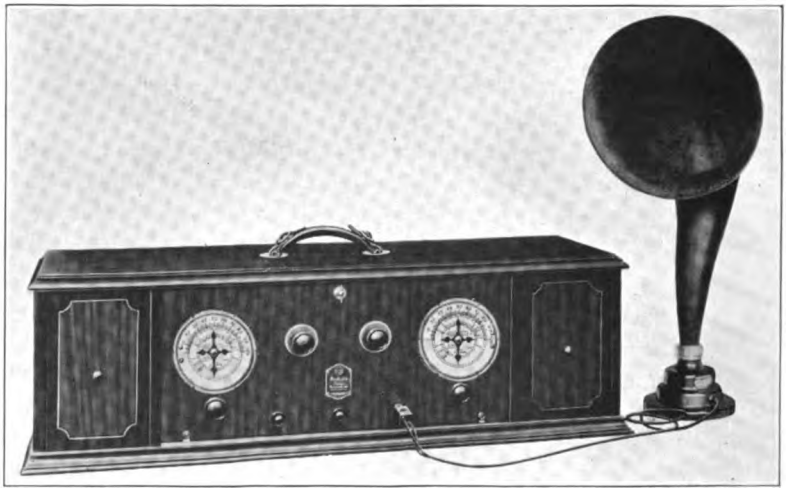

FIGURE 13

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TRANSMITTING EQUIPMENT FOR RADIO TELEPHONE BROADCASTING*

By

Edward L. Nelson

(Research Laboratories of The American Telephone and Telegraph Company and the Western Electric Company, Incorporated)

One of the most remarkable industrial developments of recent times is the advent of radio telephone broadcasting, which within the past three years has become a matter of national importance supporting a rapidly expanding industry. That a tremendous popular enthusiasm for this new art has been aroused cannot be denied. No better evidence is available than the fact that, in the face of an economic outlook which is indefinite, to say the least, some five hundred stations are now in operation, many of them representing investments of tens of thousands of dollars.

Several factors have apparently contributed toward this rapid growth. In the early stages, no doubt, the novelty of the experience was of considerable importance. The discovery of the possibilities of this latest communication development exerts a powerful stimulative influence upon the imagination which is not easily forgotten. The fact that radio presents qualities which render it ideal as a hobby has also undoubtedly aided. There are few fields in which equally satisfying results can be obtained at the expenditure of so little effort and expense while, on the other hand, consistently good performance demands a high degree of skill and more or less elaborate equipment so that interest is likely to be sustained. A more important element than either of the foregoing, however, is believed to be a keen appreciation on the part of an increasing proportion of the public of what radio is and may become as a medium of education and entertainment. The already demonstrated possibility of establishing direct personal contact with the world's most gifted minds, of hearing the greatest artists and the most renowned musical organizations, of sitting at the banquet table with the world's leaders in every field of human activity, of following athletic

*Presented before The Institute of Radio Engineers, New York, January 16, 1924. Received by the Editor, March 14, 1924.

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events of national interest play by play in distant cities, all of these things without leaving one's own fireside, constitute an appeal which even the imperfections attending the establishing of a new industry cannot destroy.

On the basis of this analysis the fundamental requirement to be met by a broadcasting equipment becomes clearly defined. It is obvious that the public in listening to the voice of the great artist desire,s above all, a natural and faithful reproduction. The slightest change in tone, any embellishment incidental to the transmission, represents a condition against which the industry must contend. Perfect imagery is the goal and future success depends upon how nearly it can be attained.

General Transmission Considerations

In general, the material to be transmitted may be divided into two classes, speech and music. There is no reason to believe, however, that the general transmission requirements for these two classes are essentially different and, in view of the important part which musical entertainment is playing and will probably continue to play in all broadcast programs, it is thought to be advantageous to define these requirements in terms of music.

Experience indicates that for the successful transmission of music, consideration must, in general, be given to frequencies up to 5,000 cycles per second and down to at least 50 cycles per second. The upper limit mentioned (5,000 cycles) lies somewhat above the range of most musical instruments, but it is nevertheless essential that the transmission band of the equipment extend to some such value. Tone quality, that characteristic by which the ear distinguishes between the note of the piano and the banjo, or between that of the violin and the flute, is a matter of overtones in varying amplitude. Naturalness in transmission, therefore, requires that these harmonics be reproduced. Further, it has been shown that the higher harmonics contribute materially to the artistic effect. Musical "brilliance" is apparently merely a matter of their presence in proper proportion.

The lower end of the frequency range is of even greater importance or perhaps it may more properly be said that it appears to be, since difficulty is usually encountered in obtaining satisfactory transmission for the lower registers. One of the most common criticisms of present day radio equipment, and probably of the phonograph as well, is that it discriminates markedly against the bass. Good reproduction is obtained for most kinds of music by utilizing a frequency range extending down to 50

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cycles. In certain types of music, however, in particular that of the pipe organ, considerably lower frequencies occur. The fundamental of certain of the percussion instruments, notably the bass drums, is also quite low.

The fact that some of the notes of certain instruments lie outside of the frequency band for which the apparatus is designed does not necessarily preclude satisfactory transmission. It is a remarkable circumstance that in certain cases the fundamental and lower overtones may be entirely suppressed without appreciable change in pitch and with very little change in quality.[1] This phenomenon is attributed to the production of subjective tones due to the non-linear characteristic of the ear. To what extent it can be depended upon to facilitate transmission of the lower frequencies is problematical. It would appear to promise little relief in so far as the percussion instruments are concerned, since in that case many of the overtones are not true harmonics, a fact which may explain the difficulty usually experienced in reproducing the drums. On the other hand, this action probably accounts for the astonishingly good performance of certain apparatus and renders it difficult to name with complete assurance the lower frequency limit required for satisfactory transmission. It will be interesting to watch the development of practices with respect to this point as the art progresses.

A second important characteristic of musical transmission is the range in volume, which for certain classes of music may represent a power change in the ratio of as great as 100,000:1. Orchestral and pipe organ numbers are particularly noteworthy in this respect. Volume is of especial importance from an engineering point of view in that it determines the load conditions under which the equipment will be operated. As broadcasting transmitters are now designed, the factor which determines their useful output range is distortion. The point at which distortion becomes objectionable under increasing load is generally quite clearly defined, being followed by a more or less abrupt change in quality. The phenomenon is usually the result of the peak alternating grid voltage in some one of the amplifier stages exceeding the polarizing voltage in the positive portion of the cycle or during the negative swing momentarily interrupting the plate current. In either case, the wave form of the transmission is modified by the peaks being flattened on one or both sides, and to the listener the music appears to break into a jumble of harsh

[1] "Physical Criteria for Determining the Pitch of a Musical Tone," H. Fletcher, "Physical Review," September, 1923.

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noise. On account of its nature, this effect has come to be known as "overloading."

It is evident that if the volume is to vary in a ratio approaching that mentioned for many musical numbers, the average will have to be set at a relatively low level in order to prevent overloading during fortissimo passages. In general, this will result in weaker signals at receiving stations and bring about a decrease in the effective range of the equipment. The practice has grown up, therefore, of manually controlling the volume by adjusting the amount of amplification employed so that while consideration is given to the demands of musical interpretation, the volume is held within considerably narrower limits, perhaps not greater than 1,000:1. While such practice admittedly precludes a really natural reproduction, experience has indicated that this discrepancy is more than compensated for by the resulting overall improvement in general transmission conditions and, as a matter of fact, the volume variation experienced with most studio music falls very nearly within this narrower range.

From another point of view, the volume range is of further importance in that it determines the permissible noise levels in the various apparatus units. At numerous points in the system slight variations in the electromotive forces of batteries or generators, mechanical vibration, induction from adjacent circuits, variable leakage paths, and similar phenomena introduce disturbances which appear as noise accompanying the transmission. It is obvious that if this noise is not to be objectionable it must be kept well in the background. In other words, its level must be low with respect to the minimum value attained by the transmission.

Apparatus Requirements

Thus far the discussion has been concerned with fundamental considerations which are in general applicable to any system for the transmission of speech and music. In proceeding to the specific requirements for a radio broadcasting equipment, it seems desirable to examine the various aspects of the problem with direct reference to the apparatus units to which they apply. For this purpose, it is convenient to divide the equipment into five groups along the following lines:

(1) Input Equipment.
(2) Radio Transmitting Equipment.
(3) Power Equipment.
(4) Control Equipment.
(5) Radio Receiving Equipment.

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The input equipment comprises the telephone transmitter or microphone, and the associated audio-frequency amplifiers.

The design of a suitable telephone transmitter for broadcasting service is admittedly one of the difficult phases of the problem. In a paper presented to the American Institute of Electrical Engineers and published in the Journal,[2] Messrs. I. W. Green and J. P. Maxfield have discussed in some detail the requirements to be met by a transmitter in a high quality transmission system and have described the condenser and double button carbon types of instruments. Both of these types have now been developed to the point where they are practically distortionless and both have found extensive application in the broadcasting field.

At the present time, the carbon transmitter is more generally employed on account of its considerably better efficiency. For broadcasting applications, however, a three-stage amplifier is usually provided for use with the carbon transmitter and an additional two stages, making a total of five, for the condenser transmitter. A second advantageous feature of the carbon transmitter is its relatively low impedance, which is of the order of 200 ohms. This characteristic makes it possible to separate the instrument from the amplifier by distances of several hundred feet without experiencing serious transmission losses or trouble due to inductive disturbances.

The advantages of the condenser transmitter are its somewhat better frequency characteristic and its freedom from carbon noise. For laboratory purposes and for the high quality transmission of the most difficult types of music it has, therefore, found some application. As has been indicated above, however, approximately five stages of amplification are subsequently required to attain ordinary telephone volume levels. The rather small capacity of the instrument (of the order of 400 micro-microfarads) also represents a more or less serious limitation to the extent of requiring cables between the transmitter and the amplifier having a very low electrostatic capacity.

It is interesting to note that in popular discussion the "microphone" is frequently regarded as the prime offender in case of poor transmission, and as the secret of success if satisfactory results are obtained. It is significant, however, that before modulation actually occurs, the output of the telephone transmitter or microphone is often passed thru as many as seven stages

[2] "Public Address Systems," I. W. Green and J. P. Maxfield, "Journal of the American Institute of Electrical Engineers," volume XLII, number 4, April, 1923.

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of amplification in which the power is increased several million times. That such a feat is possible at all is a glowing tribute to present day amplifier technique, and this fact should be given due weight when passing judgment on the "microphone."

The principal requirements to be met by the amplifier associated with the telephone transmitter in the input equipment are those relating to frequency range, volume range, and noise level, which have already been discussed. A further matter of some importance is that concerned with amplification control. The necessity of providing for manual variation in amplification to reduce the range in volume in the higher powered stages has already been mentioned.

The radio transmitting equipment must afford means of generating alternating currents of a frequency corresponding to the operating wave length, means for modulating these currents faithfully in accordance with the desired transmission, and suitable radiating means, in the form of an antenna.

The vacuum tube oscillator in its present form is a very satisfactory solution to the problem of generation. Since a broadcasting station usually operates on a single assigned wave length, a relatively simple circuit arrangement can be employed. Under the present regulations, however, all coastal stations must be capable of transmitting for ship emergency service on wave lengths of 300 and 600 meters. The equipment must, therefore, admit of adjustment on occasion to these wave lengths.

Other requirements arising from the provisions of the radio regulations relate to frequency stability and to freedom from harmonics. Frequency stability is an almost essential pre-requisite to the success of any scheme looking to the intensive utilization of the ether such as has been put into effect within the past year. If stations are to be successfully operated with frequency intervals between adjacent wave lengths as small as 10,000 cycles, a high degree of constancy in this respect must obviously be maintained, not only during the course of any individual program, but also from day to day and from month to month. For those who have not had the privilege of listening thruout an entire evening to the beat-note from a distant transmitter superimposed upon the program of their favorite broadcasting station, it may be stated that if proper precautions are taken no difficulty is ordinarily experienced in holding the frequency constant within considerably narrower limits than those specified by the present regulations, that is $\pm 2{,}000$ cycles. Since common use is made, however, of transmitters in which the antenna forms part of the

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oscillatory circuit or is coupled thereto, it is evident that such performance requires the antenna to be so secured that even the effect of a high wind will not bring about a material change in its constants.

With respect to freedom from harmonics, the present regulations covering broadcasting stations merely contain the provision that the transmitted wave shall be free from harmonics. Unfortunately, this requirement is very nearly an exact statement of the case. The sensitivity of modern receiving equipment and the extremely small amount of power necessary to produce an audible beat-note require the suppression of harmonics to a most unusual degree and render the problem exceedingly difficult. Apparently satisfactory performance is now being attained, however, thru the use of properly designed coupled circuits.

Primarily, the system of modulation employed in a broadcasting transmitter must be one which will enable it to meet the rigorous transmission requirements. On account of its excellence in this respect, as well as its simplicity, the arrangement commonly known as "constant current" or "plate current" modulation has found extensive application. In this system, the amplified transmission is superimposed upon the direct voltage impressed upon the plate of the oscillator tubes. Under these conditions, the output voltage of the latter varies proportionally, producing the desired modulated radio-frequency current in the antenna circuit. The operation of such a system has been discussed at length in a paper presented to THE INSTITUTE OF RADIO ENGINEERS in 1921 by Mr. R. A. Heising,[3] and need not be reviewed here.

The power equipment for a radio broadcasting station is subject to no requirements of an unusual nature with the possible exception of freedom from electrical noise. This factor is of no little importance, however, and exercises considerable influence on the choice of power sources for use in the various circuits.

The effect of slight voltage variations in producing disturbances which appear to the listener as noise has already been mentioned. In general, it is essential that all power sources be substantially constant in voltage. The absolute value of the allowable voltage fluctuation is variable, depending upon the position of the battery or generator in the circuit. If associated with one of the early stages where the transmission level is low and where any fluctuation will be subsequently amplified to a considerable extent, relatively rigid requirements are imposed and batteries

[3] "Modulation in Radio Telephony," R. A. Heising, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 9, number 4, August, 1921.

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may be required. If, on the other hand, the source supplies power to one of the later stages, relatively large fluctuations in voltage may occur without producing effects comparable in magnitude with the desired transmission, and rotating machines or rectifiers may prove to be entirely satisfactory. The power supply to the plates must, in general, meet more rigorous requirements than that for the filaments. If, however, grid biasing voltages are obtained by the use of resistances in the filament circuit, a rather common practice, the current in that circuit must, for obvious reasons, be maintained substantially constant.

When the power sources themselves are not inherently free from noise to the required degree, the most direct solution lies in the use of suitable electrical filters. Considerable improvement can generally be effected in this manner, but such combinations are not always practical possibilities on account of the cost and bulk of large condensers to withstand high voltages and of large inductances to carry relatively heavy direct current.

Another requirement of some importance applying particularly to power sources in the filament circuits, is that relating to voltage regulation. The life of a vacuum tube is very definitely limited by that of the filament. The latter slowly evaporates and in time burns out. This process is necessarily a function of the filament current and it is an experimental fact that an increase in the current of a few percent above the normal value will reduce the life to one-half or even less, depending upon the type of filament, the design of tube, and the service which it is called upon to give. Close voltage regulation is, therefore, essential and is one of the factors to which particular attention must be given in preparing specifications for filament machines.

A third requirement of considerable commercial importance is that of meeting the power supply available at the point of installation. For a number of reasons the majority of broadcasting stations now in service are building top stations. Many of them are operated by newspapers and stores having isolated power plants, or are located in business districts where alternating current power is not readily obtainable. Under these conditions, the application of rectifiers as a source of high voltage power is at an obvious disadvantage, and direct current generators appear to be the most satisfactory solution.

In broadcasting, to an even greater extent perhaps than in other amusement enterprises, the program must go forward smoothly and briskly to obtain the most favorable reaction from the audience. Delays of even a relatively few seconds often be-

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come painfully evident since the sense of hearing alone is involved, and as soon as the transmission is interrupted the listener is beyond the influence of the personality of the entertainer. It is of primary importance, therefore, that such facilities as are required to expedite the program to the utmost should be provided.

The control equipment required depends largely upon the class of service in which the station operates. The smaller stations necessarily serve a limited audience and often employ considerable mechanically reproduced music. Under these conditions, the simplest possible arrangement will generally suffice. On the other hand, many of the larger stations now number their audiences in tens of thousands, and the talent available often includes figures of national or international prominence. Two or more studios are frequently employed, and the use of wire line connections to outlying points is finding increasing application. Under such circumstances, a well-worked-out control system is essential.

In general, the facilities required may be classified into three separate groups. In a broadcasting station, it is just as important to close and open the circuit promptly at the proper time as it is to raise and lower the curtain in the theatre. There is also the problem of insuring that nothing but the desired transmission is broadcast, since failure in this respect undoubtedly detracts materially from the artistic effect produced and often results in considerable embarrassment to the station personnel or to the talent. These conditions require that control over the outgoing transmission be vested in the studio director or announcer and that suitable provision be made in the studio for this purpose. The apparatus employed, in addition to the necessary keys and relays for circuit control, must include suitable indicating means such as a series of signal lamps, so that circuit conditions at any particular instant will be evident to all persons involved.

Another requirement arises from the close cooperation which must be maintained between the studio and operating personnel and the fact that in many stations the studio and operating room are widely separated. A suitable intercommunicating telephone and signal system is almost an absolute essential. Both audible and visible signals are desirable, and suitable means must be provided to prevent such signals being broadcast.

A third requirement relates to monitoring facilities. The operator of a radio broadcasting station has considerably more

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important functions than those of telegraphist, "trouble shooter," and wireman. Not only is he the critic upon whom the studio director in most cases must depend for information concerning the effect which the audience is actually observing, but also thru his control of the input volume he himself contributes materially to the excellence of the program or tends to spoil it entirely. It is quite necessary, therefore, that he be provided with suitable apparatus facilities to enable him to obtain the desired results. Preferably, these should take the form of a suitable loud speaker equipment arranged for monitoring the output of the speech amplifier and, by means of a suitable rectifier, the radio frequency output of the transmitter.

In so far as receiving apparatus is concerned the requirements to be met are, at best, somewhat indefinite. Since a large number of broadcasting stations may be classified as "coastal stations," however, equipment suitable for maintaining the receiver watch required by law is essential. In this connection, the original practice, it will be remembered, involved an interruption of the program every fifteen minutes for three minutes to "listen in." Such a procedure was necessarily of short duration and in almost all stations facilities are now provided to permit of reception at 500 kilocycles (600 meters) while the local transmitter is in operation. This requires a separate receiving antenna and imposes certain restrictions upon the design of the receiving apparatus. In addition to its use in the reception of telegraph signals, it is also desirable that the receiver be suitable for employment in the broadcasting range and be capable of the high quality reproduction of speech and music, since, on occasion it may be employed to considerable advantage to supplement the regular monitoring facilities and to observe the programs from other stations.

The Western Electric 101-B Radio Broadcasting Equipment

Having discussed various aspects of the general design problem, a brief description of a commercial system of the type which has found application in some forty of the larger broadcasting stations in the country may be of interest.

The transmitter in its housing is, no doubt, already a familiar object since it has appeared with great frequency in published photographs of broadcasting activities. An idea of the arrangement in more detail, however, may be obtained from Figure 1. On the left is shown the housing with the cover removed. The

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transmitter is supported by means of a series of coiled springs to reduce noise effects due to mechanical vibration. A 12-foot (3.7 m.) flexible cord is ordinarily provided which terminates in a standard three-wire attachment cap. When in service, the latter is inserted in a suitable receptacle usually mounted on the baseboard in the studio.

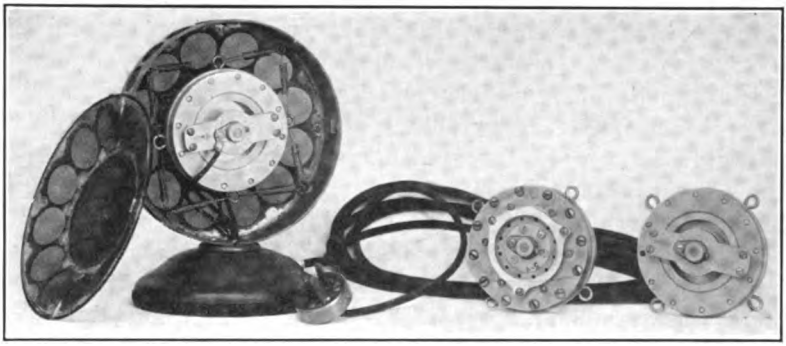

FIGURE 1—Carbon Transmitter and Housing

The appearance of the transmitter itself from the rear, and from the front, respectively, is shown by the two views on the right. A light, thin, metallic diaphragm is employed which is securely clamped between the two heavy exterior rings. An interior ring stresses the diaphragm like a drumhead so that its natural frequency is brought above the range which it is desired to transmit. In this manner "blasting" effects due to diaphragm resonance are practically eliminated. Its performance in this respect is still further improved by the application of air damping. A perforated metal plate is mounted close to the diaphragm in the rear, providing a thin film of air which serves to increase the damping and bring about a much flatter frequency characteristic. Two carbon buttons are employed, one on each side of the diaphragm. One of these is mounted in the center of the damping plate; the other is supported by means of a heavy bridge secured to the clamping rings. It will be noted that, with this arrangement, as the diaphragm vibrates, the pressure on one button is released when that on the other is increased. For this reason it is sometimes referred to as the "push-pull" type. These various improvements have resulted in a transmitter which permits of a reproduction of speech and music not greatly different from the original. As has been indicated, however, two or three stages of amplification are required to attain ordinary telephone volume levels.

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The vibratory system of the condenser transmitter is similar in construction to that of the carbon unit. The buttons and bridge are omitted, however, and the rear damping plate is carefully insulated to serve as the second plate of the condenser. The capacity of the condenser thus formed is approximately 400 micro-microfarads and the variation brought about by the vibration of the diaphragm is of the order of one-hundredth of one percent. As has been stated, this device in its present form is less efficient than the carbon transmitter and requires an additional two-stage amplifier if equal volume is to be obtained. Also, it requires special handling in certain respects. On account of its low capacity and correspondingly high impedance, a relatively short cable of low electrostatic capacity must be employed to connect with the initial stage of the amplifier if serious transmission losses are not to be experienced. In addition, its high impedance results in the amplifier being operated with the input of the first amplifier tube practically on open circuit and the latter becomes sensitive to inductive disturbances which introduce noise into the circuit. The fact that four or five stages of amplification is required to attain ordinary telephone volume levels is, in itself, a disadvantage from a noise point of view, since under these conditions, the problem of adequately controlling tube noises, variable leakage paths, battery voltage irregularities, and the like, becomes difficult. On the other hand, the condenser transmitter is the most highly developed device of its kind now available; when properly maintained it is free from the background of noise which is sometimes so objectionable with the carbon instrument, and it affords somewhat better transmission than the latter. Where the very best results are desired, therefore, irrespective of expense, it finds application.

A type of condenser transmitter equipment designed primarily for studio use is shown in Figure 2. The transmitter is mounted in a housing of the usual type and fitted with a special 6-foot (1.8 m.) low capacity cord. The amplifier unit takes the form of a cabinet of the phonograph type with the amplifier proper mounted in the upper portion on a horizontal panel. Adequate shielding is afforded by a metal backing on the panel and a removable metal cover attached thereto. Space is provided in the lower portion for a 6-volt storage battery for heating the filaments and a 200-volt dry battery for the transmitter and plate circuits. The amplifier circuit is shown schematically in Figure 3, which probably requires little explanation. Provision is made for controlling the device from the radio room by means

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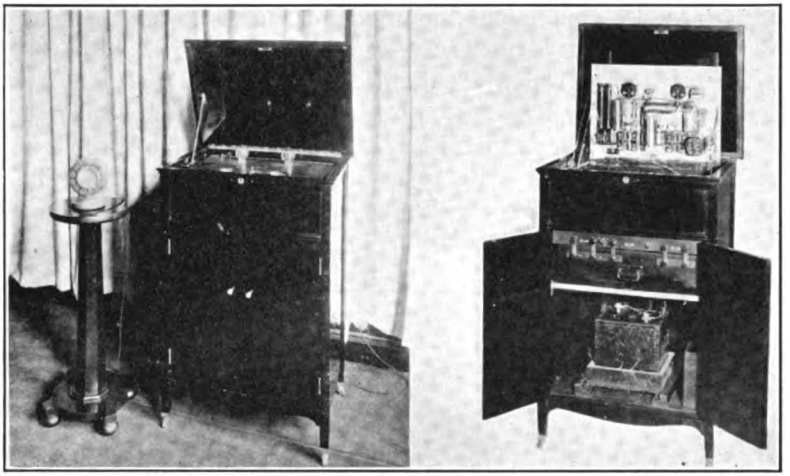

FIGURE 2—Condenser Transmitter Equipment

of a relay in the filament circuit operated by the current which would normally flow thru the buttons of the carbon transmitter. Since the amplifier is fitted with an output cord similar to that used with the carbon transmitter, it may be plugged into the circuit interchangeably with the latter and employed without change in the operating procedure.

The studio signaling and control unit is shown in Figure 4. As is evident, it takes the form of a wooden cabinet similar in outline to the ordinary subscriber's set, having the necessary

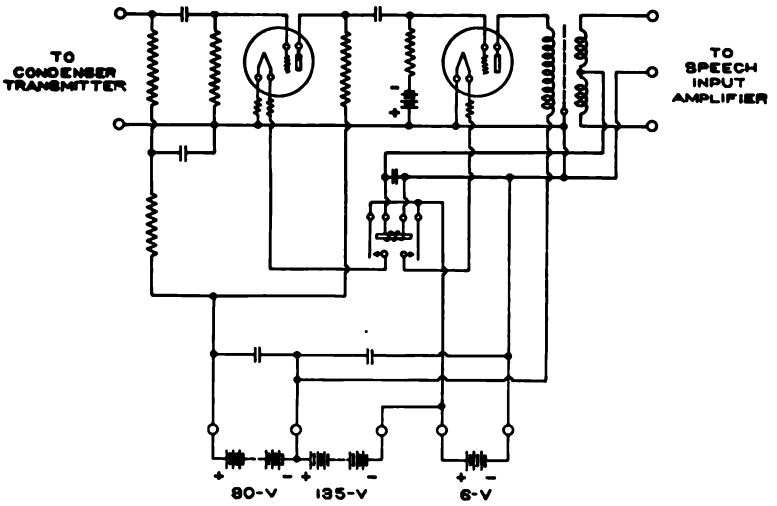

FIGURE 3—Schematic of Amplifier for Condenser Transmitter

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switches and signal lamps mounted on the front panel. The telephone handset permits of direct communication with the operating room. The white lamp, key and push button to be seen on the left and at the bottom of the panel are associated with a signaling buzzer circuit. The lamp at the right, which is green, is controlled from the operating room and serves to indicate to the studio director when the radio equipment is in operation and ready for the program to begin. The key immediately

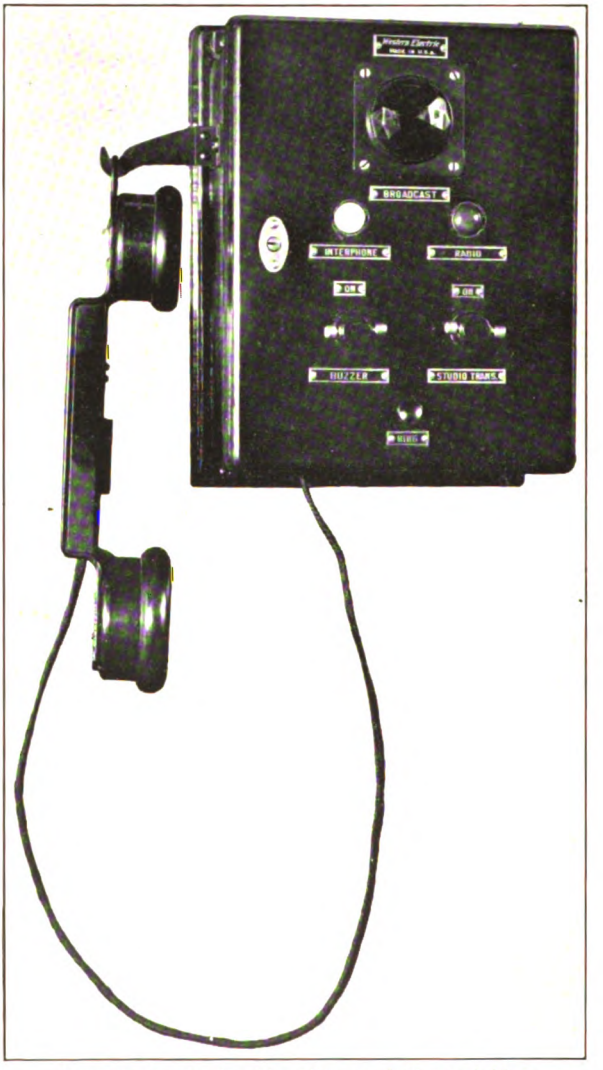

FIGURE 4—Studio Signaling and Control Unit

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below it serves to control the studio transmitter circuit and, in addition, when thrown to the broadcast position, opens the circuit of the signaling buzzer and lights a lamp behind the large red bull's-eye at the top of the unit.

The speech input equipment is shown in Figure 5. As will be

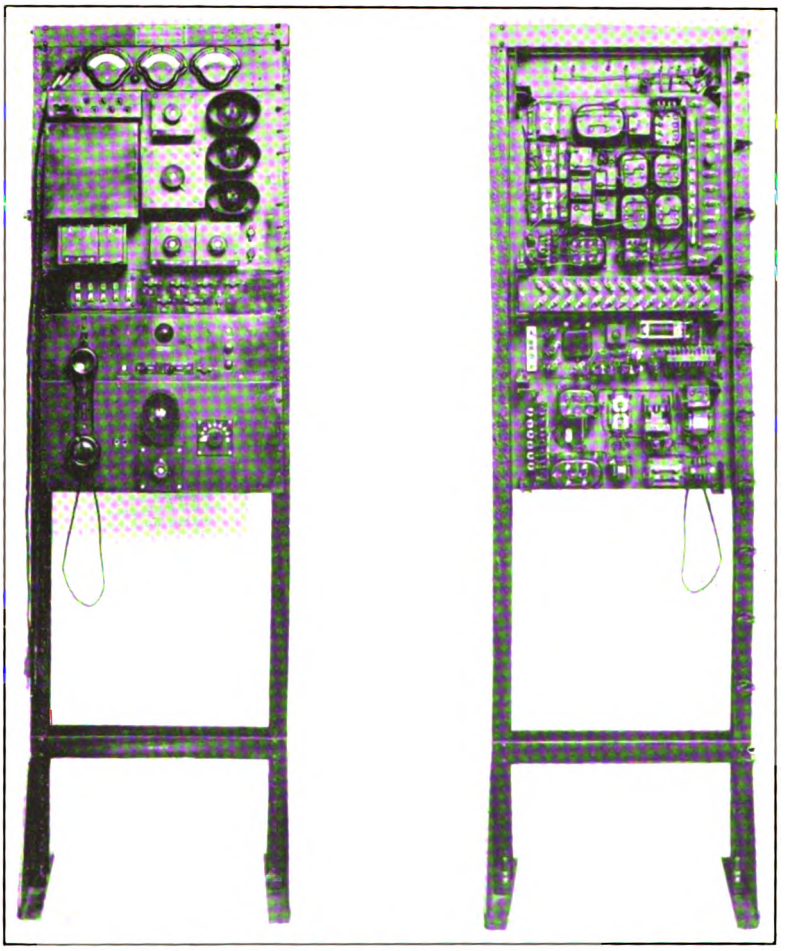

FIGURE 5—1-B Speech Input Equipment

noted, this apparatus consists of five separate units: a meter panel, a three-stage amplifier, a switch and fuse panel, a signal and control panel, and a single stage monitoring amplifier. In the rear view, the metal covers, which normally enclose the individual units, have been removed to enable the apparatus to be seen in more detail.

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The circuit of the three-stage amplifier is shown schematically in Figure 6. Vacuum tubes having an amplification constant of

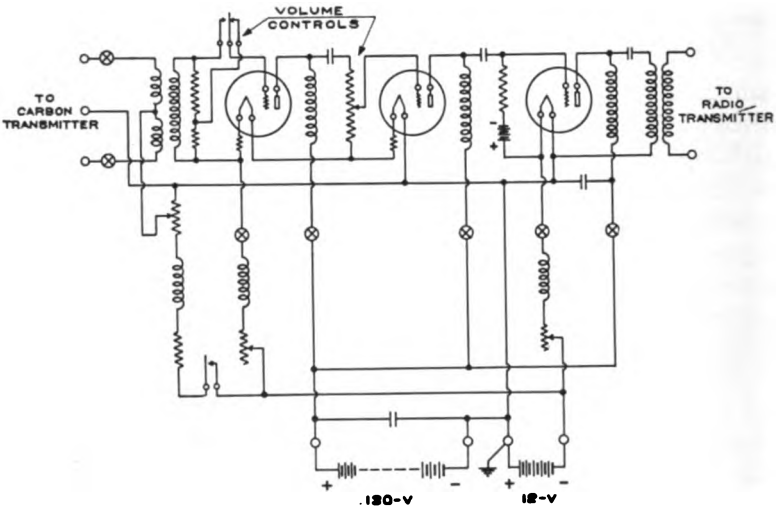

FIGURE 6—Schematic Diagram of Speech Input Amplifier

approximately thirty are employed in the first two stages. The last stage employs a tube having a considerably lower amplification constant but a greater power capacity. Potentiometers in both the first and second stages afford a wide range in amplification control. Power for the filaments is obtained from a suitable 12-volt storage battery. A 130-volt battery consisting of standard Number 6 dry cells supplies power to the plate circuits. Grid voltages for the first two tubes are obtained by utilizing the voltage drop over suitable resistances in the filament circuit. A grid battery of the flashlight type is employed with the third tube. The maximum amplification to be had with this amplifier is 85 miles[4] representing a power ratio of 1.2×10^8. By means of the potentiometers, the amplification can be varied by 3-mile steps from the maximum down to approximately 25 miles.

The signal and control panel is employed in connection with the studio unit, which has already been described. On this panel, in addition to the intercommunicating telephone circuit and the buzzer and lamp signaling facilities, arrangements are made for the control of the radio transmitter, the microphone circuits, and the monitoring circuits. The radio transmitter control feature

[4]Miles gain $= \frac{21.13}{2} log_{10} \frac{P_2}{P_1}$ where P_1 is the input power, and P_2 the output power.

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takes the form of a key which operates a relay in the field circuit of the plate generator. In the microphone circuit, a suitable three-position, multiple-contact key is provided enabling the studio transmitter, the operator's transmitter, or a third instrument (or a wire line) to be connected to the input terminals of the amplifier, as the occasion may require. These two keys are electrically interlocked with the green "RADIO" lamp on the studio unit, so that both keys must be thrown in the proper direction before a signal is given to the announcer. For monitoring purposes, a suitable key is provided to connect the monitoring loud speaker to either a special vacuum tube rectifier which is built into the radio transmitter and coupled to the antenna circuit, or the monitoring amplifier which is connected permanently to the output terminals of the three-stage amplifier. By the operation of this key it is possible, therefore, for the operator to listen either to the outgoing transmission or to the input to the radio transmitter and to keep an accurate check on the operation of the equipment. When the microphone control key is thrown to the operating room position, a red lamp on the panel is lighted and the monitoring loud speaker is disconnected, irrespective of the position of the monitoring key, to prevent "singing" between the operator's transmitter and the loud speaker. A white guard lamp is also provided on the panel to indicate to the operator when the studio control key is thrown to the broadcast position.

The monitoring amplifier is a single stage unit of the usual type but with transformers designed for high quality transmission. A potentiometer is provided for volume control. A suitable loud speaker mounted on the operator's desk is associated with this amplifier for monitoring purposes.

Front and rear views of the radio transmitter are shown in Figure 7 and the circuit employed is represented schematically in Figure 8. This unit is of the oscillator-modulator type employing two 250-watt tubes as oscillators and two similar tubes as modulators. One 50-watt tube is also utilized as a speech amplifier working into the modulator tubes. The oscillator circuit is of the Meissner type, that is, both the grid and plate circuits are inductively coupled to the oscillatory circuit, but it departs from the usual arrangement in that an adjustable condenser is employed in parallel with the plate coil. This variation has been shown to afford a simple and convenient means of adjusting impedance conditions in the circuit so as to bring about efficient operation. As will be noted, the oscillator does not work directly into the antenna, but a closed oscillatory circuit

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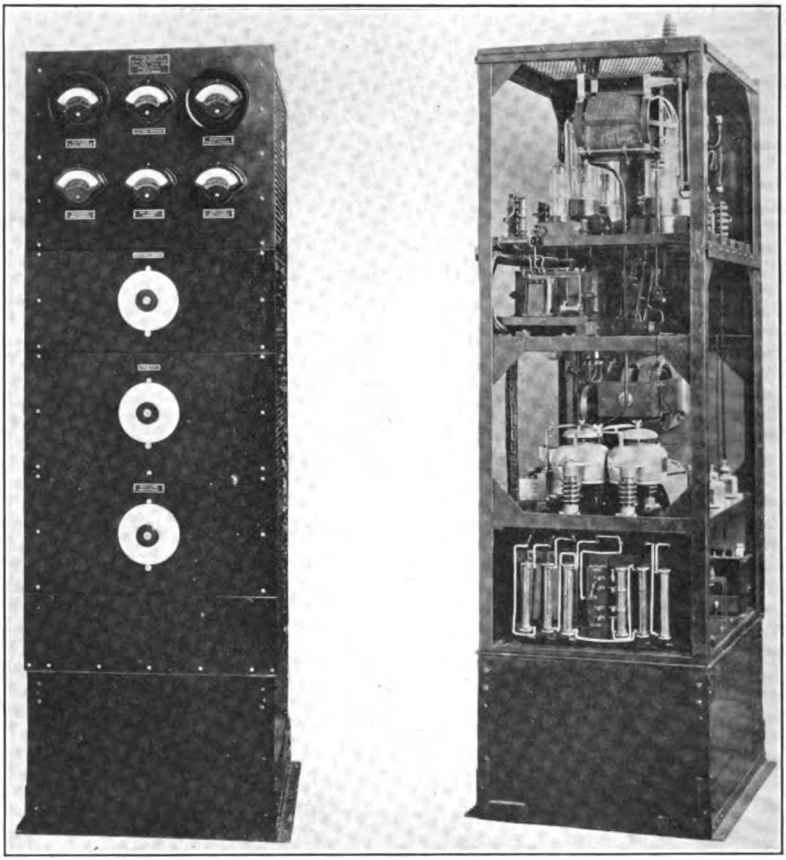

FIGURE 7—1-B Radio Transmitter

is employed which is coupled to the antenna circuit by means of a mutual capacity. This arrangement has been adopted primarily for the suppression of harmonics. The use of two tuned circuits and capacity coupling affords a high degree of discrimination against these higher frequencies, and radiation is confined almost entirely to the desired frequency band. The use of coupled circuits also results in somewhat better frequency stability than that representative of a transmitter employing the antenna as a part of the oscillator circuit, but dependence in this respect must be placed primarily in a rigid antenna structure.

Another feature of possible interest in connection with the radio frequency circuits is the antenna discharge resistor which appears on the schematic and may be seen in Figure 7 in the upper right-hand corner of the rear view. Experience has indi-

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cated that with many antennas, if a series condenser is employed, thus interrupting the direct current path to ground, static charges will accumulate which may build up a voltage sufficient to break down the safety gap of the condenser. A number of stations have reported occasional interruptions arising from trouble of this nature. Observations to date indicate, however, that a relatively high resistance connected as shown is a complete solution to the problem.

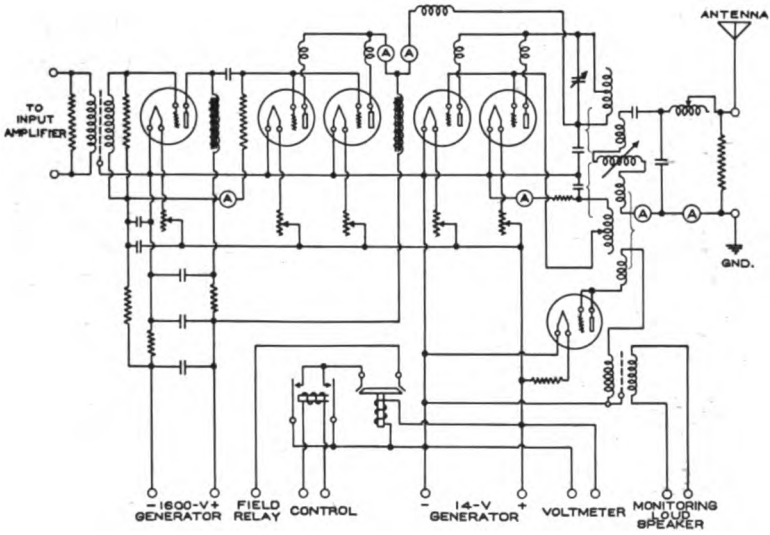

FIGURE 8—Schematic of 1-B Radio Transmitter

The audio frequency portion of the transmitter circuit, embracing the speech amplifier and the modulator, is essentially a two-stage, audio frequency amplifier. The output circuit of the last stage of this amplifier (the modulator) includes the plate circuit of the oscillator. The power supply to the latter may be considered to consist, therefore, of a direct component upon which is superimposed an audio frequency component corresponding to the desired transmission. Under these conditions the output of the oscillator varies in proportion and the envelope of the time-antenna current curve is a faithful reproduction of the signal voltage impressed upon the plate circuit. It is interesting to note in this connection that, within the usual operating limits, the distortion which can be traced to the radio frequency portion of the transmitter is negligible. In other words, by suitable design of the radio frequency circuits, the matter of insuring high quality

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transmission may be resolved into proper treatment of the audio frequency circuits, and all of the accumulated experience of the telephone art brought to bear upon the problem.

A somewhat novel feature of the audio frequency circuit is the method employed for obtaining negative grid voltages for the speech amplifier and modulator tubes. A resistance is employed between the negative terminal of the plate generator and the filaments as shown in Figure 8. The grid circuit is connected through a filtering network to such a point on this resistance that the IR drop due to the plate current represents the required negative grid voltage. This method renders a grid battery unnecessary, an important consideration from a maintenance point of view.

The mechanical features of the transmitter are clearly shown in the photographs and require little explanation. The meters and controls are centralized upon the front panel. Meters are provided in the antenna and closed oscillatory circuits, in the oscillator plate and grid circuits, and in the modulator plate and grid circuits. When the equipment is in operation, the needle of the modulator plate meter swings at syllable frequency and affords a convenient, altho somewhat rough, means of adjusting the input amplifier to give the proper volume level. The function of the meter in the grid circuit of the modulator is to serve as an "overload" indicator. Overloading has already been mentioned and its causes briefly discussed. In the equipment as designed, it is occasioned by the peak signal voltage impressed upon the modulator grid exceeding the polarizing voltage in the positive direction. When this occurs, a momentary current will flow from the filament to the grid and the meter will give an indication. The operator is, therefore, afforded a positive means of detecting overloading and a definite operating procedure may be worked out to insure best results.

The three dials shown control the antenna tuning, the frequency, and the oscillator adjustment, respectively. The first is associated with the antenna tuning coil shown in the top of the set behind the tubes. The second operates upon the oscillator coil, which appears near the center of the rear view and serves to vary the inductance of the winding in the oscillatory circuit. The oscillator adjustment control is connected with a variable air condenser joined in parallel with the plate coil as previously described.

The condensers in the radio frequency circuits appear below the oscillator coil. The audio frequency coils and condensers are

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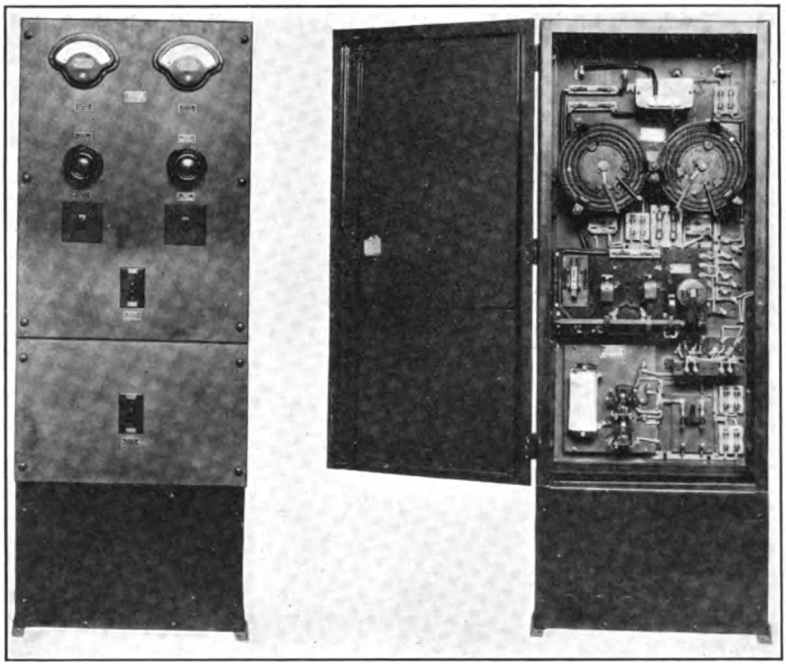

FIGURE 9—Power Panel

mounted under the tube compartment and in the bottom of the unit behind the resistance unit assembly.

The power apparatus for the radio transmitter is shown in Figure 9 and Figure 10. The motor-generator consists of three units mounted on a fabricated base: a suitable motor, a 1,600-volt 1.25-ampere generator, and a 16-volt 30-ampere generator. The high voltage generator is constructed with two 800-volt armature windings and employs two commutators. Current for

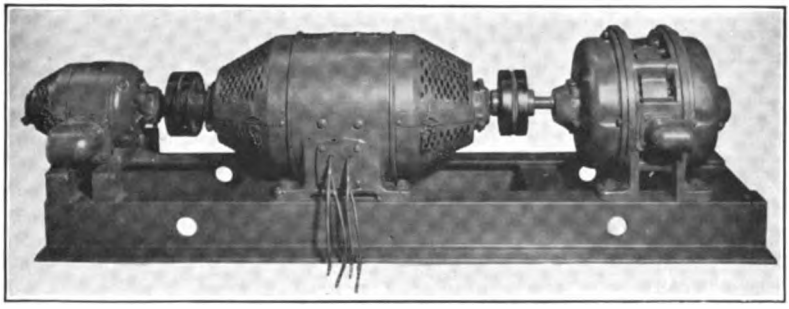

FIGURE 10—Motor Generator

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the field is obtained from the low voltage unit, which is self-excited.

The power panel is of the cabinet type. It includes a suitable motor starter which is controlled from a push-button station on the panel and on the operator's desk, field rheostats and voltmeters for each generator, an electromagnetically operated switch in the plate circuit (which also serves as a circuit breaker), and rotary snap switches in the filament and high voltage field circuits. No live metal parts are exposed on the panel and, as further protection to the operator, a door switch is provided which prevents the motor from being started until the door of the cabinet is closed.

The radio receiver is shown in Figure 11. This unit consists

FIGURE 11—2-C Radio Receiver

of a detector and two stages of audio frequency amplification utilizing tubes similar to the VT-1. Coupled radio frequency circuits are employed without regeneration. Suitable jacks are provided, permitting the detector to be used alone or in combination with one or both of the amplifier stages as required. Power for the filament and plate circuits is obtained from the batteries that supply the speech input equipment.

In discussing the individual apparatus units, an attempt has been made to point out in each case the relationship to the other parts of the equipment. The description would hardly be complete, however, without some further indication of the nature of the system as a whole. In Figure 12, therefore, the general arrangement is represented diagrammatically. All of the units are shown and the various connections between them, the broadcasting circuits, the signal and control circuits, and the power circuits, are indicated by the different types of lines. Viewed in

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combination with the detailed descriptions previously given, this drawing will enable the general aspects of the system to be readily understood.

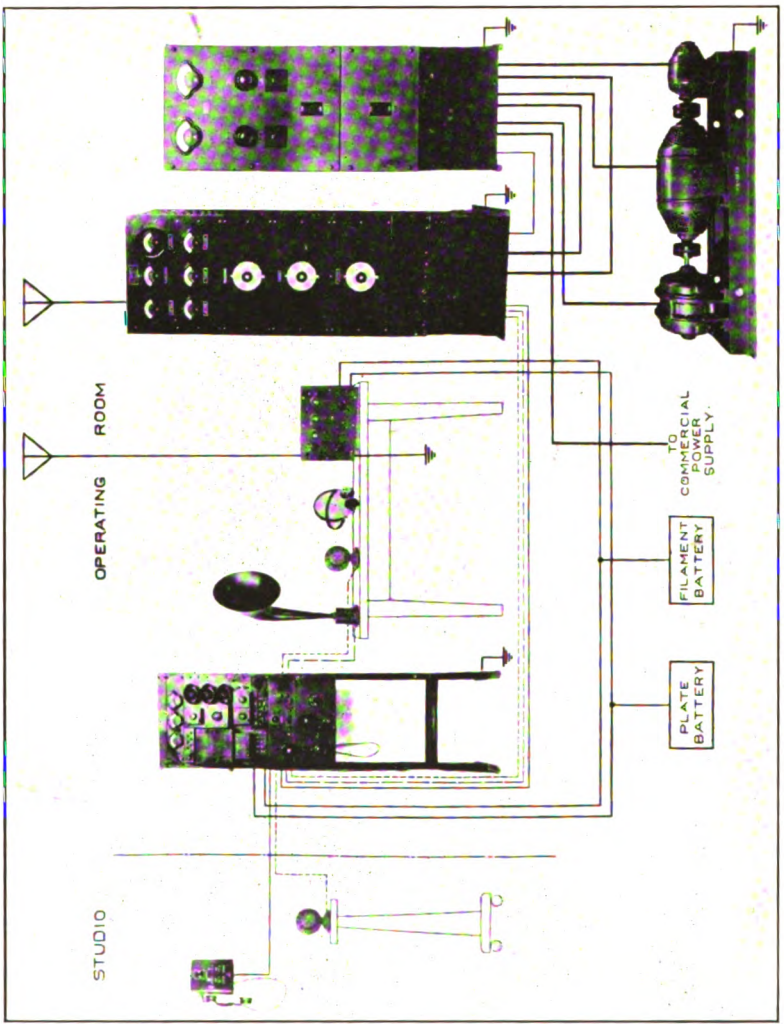

FIGURE 12—101-B Radio Telephone Broadcasting Equipment

On account of the severe transmission requirements imposed upon this class of apparatus, the performance capabilities of the equipment, especially its frequency-transmission characteristic, will probably be of considerable interest. The characteristic for the equipment when employing the carbon transmitter is shown

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in Figure 13. This curve is a composite based upon single frequency measurements made on the individual apparatus units and may be considered to represent the relative magnitudes at the various frequencies of the effective alternating component

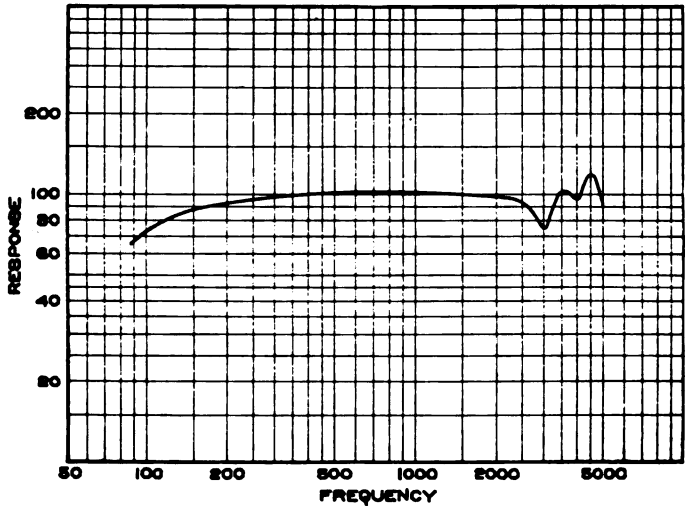

FIGURE 13—Frequency Response Characteristic with Carbon Transmitter

in a suitable rectifier circuit coupled to the antenna when the sound pressure impressed upon the diaphragm of the transmitter is maintained constant. In other words, if the effects of standing waves and other similar acoustical phenomena are eliminated, and if a pure tone of constant intensity but variable pitch is produced in the studio, this curve represents the relative intensity of the sound which might be expected from a distortionless receiving device. Logarithmic scales are employed for both ordinates and abscissas since there is considerable evidence to indicate that the response of the ear to variations in both intensity and pitch follows a logarithmic law. As will be noted, substantially uniform transmission efficiency is indicated except at the extreme ends of the frequency range where some slight deficiencies become evident. These can only be detected by careful measurement, however, and it is doubtful if their effect upon transmitted music would be apparent even to the artistically trained ear.

CONCLUSION

The general considerations involved in the design of a system for the high quality transmission of speech and music have been men-

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tioned briefly, and the requirements to be met by the various apparatus units in a radio telephone broadcasting equipment have been discussed in some detail. A commercial 500-watt transmitting equipment has been described and its performance capabilities illustrated. It has been shown that apparatus can be produced commercially which will afford transmission at substantially uniform efficiency throughout the entire frequency range commonly employed in music.

From a quality point of view, at least, therefore, transmitting equipment for broadcasting applications may be said to have already reached an advanced stage of development. A standard of performance has been attained which it is not believed to be presumptuous to claim is an artistic, as well as a technical success. The next few years will, no doubt, bring forth contributions of great value, new and more efficient telephone transmitters, better vacuum tubes, more stable circuits, and less expensive apparatus, but in view of present attainments, radical improvement in the quality of the transmission can hardly be expected. If further improvement in quality is desired by the listener, therefore, it must be obtained elsewhere than in the transmitting equipment.

New York, N. Y.
January 16, 1924.

SUMMARY: The general transmission considerations applying to any system for the high quality transmission of speech or music are outlined briefly, and the specific requirements to be met by the various apparatus units in a radio broadcasting equipment are discussed in some detail. The standard Western Electric 500-watt broadcasting equipment, which has found application in some fifty of the larger stations in this country and abroad, is described. Its performance capabilities are illustrated and it is indicated that a standard of performance has been attained which renders possible reproductions not substantially different from the original.

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AN ANALYSIS OF TWO TRIODE CIRCUITS*

By

John H. Morecroft

(Trowbridge Fellow, Columbia University)

and

Axel G. Jensen

(Scandinavian-American Fellow, Columbia University
New York City)

There seems to have been but little exact analysis of triode circuits, with the necessary experimental verification, without which the theoretical work is not complete. It is with the idea of helping to fill this gap that the following analyses and experiments are reported.

The plate current of a triode can be written

$$I_p = a\,(E_p + \mu E_g)^x$$

in which the symbols have their usual meaning. The exponent x is different in different types of tubes, and varies greatly in a given tube as the voltages used in the grid and plate are changed from small to large values. For the tube used in the following tests the curve represented by the above equation is given in Figure 1. The grid was held at zero potential and plate voltage changed as indicated.

The input circuit of a tube, grid-to-filament, has both conductance and capacitance. The former may be made negligible by negative bias and the latter is negligible at low frequencies. Such was the case in our tests.

The output circuit, plate-to-filament, also has resistance and capacitance; the latter was negligible in our tests, but of course the plate-filament resistance can seldom be neglected, and it was not, in our analyses and experiments. The resistance of the plate-filament circuit of the tubes has three different values, depending upon how the resistance is defined.

(a) Continuous current resistance, R_o. From the plate current equation $R_o = \dfrac{E_p}{I_p} = \dfrac{1}{a\,(E_p + \mu E_g)^{x-1}}$; or, if E_g zero,

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$$R_o = \frac{1}{a E_p^{x-1}} \tag{1}$$

(b) Alternating current resistance, plate voltage alone varying.

$$R_p = \frac{\Delta E_p}{\Delta I_p} = \frac{\partial E_p}{\partial I_p} = \frac{1}{a x (E_p + \mu E_g)^{x-1}} = \frac{R_o}{x} \tag{2}$$

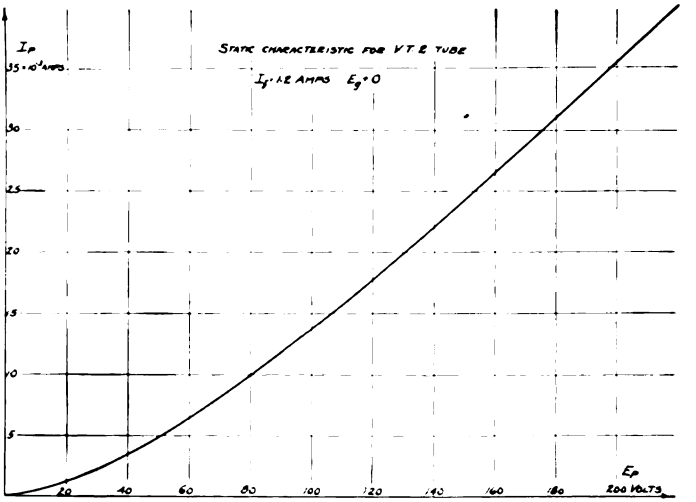

FIGURE 1

This resistance is constant only for small values of ΔE_p. For larger increments, the resistance must be written as a constant plus a series of harmonic terms. The value of the constant term increases with amplitude of ΔE_p, approaching R_o as a limit.

(c) Alternating current resistance, plate and grid voltages both varying.

$$d I_p = a x (d E_p + \mu d E_g)(E_p + \mu E_g)^{x-1}$$

$$R_p' = \frac{d E_p}{d I_p} = \frac{1}{a x (E_p + \mu E_g)^{x-1}} - \mu \frac{d E_g}{d I_p}$$

$$= R_p - \mu \frac{d E_g}{d I_p} \tag{3}$$

Generally, E_g is varied by variations of I_p and it may be that the value of $\frac{d E_g}{d I_p}$ is either positive or negative, according to the manner in which E_g is made to change by changes in I_p. In case $\frac{d E_g}{d I_p}$ is positive, and sufficiently large, R_p' may be negative, that

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is, altho the change in plate voltage $d E_p$ is positive, the accompanying change in plate current is negative.

In the following analyses, by the term "plate resistance," the value of R_p (case *b*) is always intended. This value is always positive and, in magnitude, is somewhat less than the "B" battery voltage divided by the steady value (average value) of the current furnished to the tube by this battery.

Oscillatory Circuit Connected to Plate

The first circuit to be considered is that given in Figure 2.

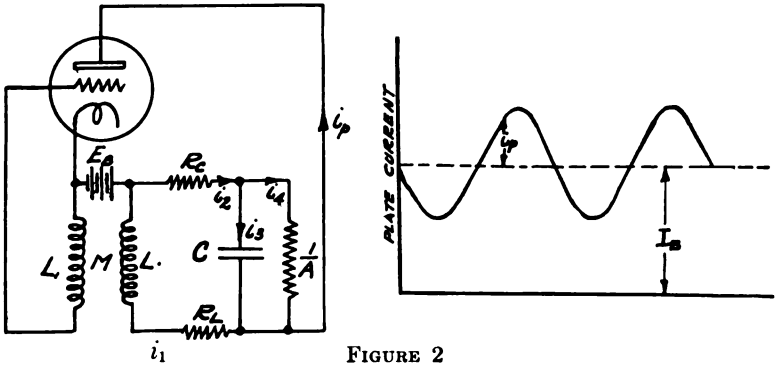

FIGURE 2

A small increment in plate voltage or grid voltage will produce currents in the circuit as indicated. It is supposed that the condenser has leakage, $\frac{1}{A}$, (conductance $= A$), and that the resistance of the condenser branch is R_c. For a good condenser, both of these may be neglected, but if the condenser happens to be an antenna, such is not possible. In the following equations e_p is the instantaneous value of the alternating component of the plate voltage, e_g is the instantaneous value of the alternating grid voltage, i_p is the instantaneous value of the alternating component of the plate current. In the analysis to follow, we assume the grid takes a negligible current; at low frequencies the charging current is negligible and by suitable bias the conduction current may also be made negligible. Then

$$i_p = \frac{e_p + \mu e_g}{R_p} \text{ or } i_p R_p = e_p + \mu e_g \tag{4}$$

$$e_g = -M \frac{d i_1}{d t} \text{ (This defines polarity of } M) \tag{5}$$

$$i_p = i_1 + i_2 \tag{6}$$

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$$e_p = -i_1 R_L - L\frac{d\,i_1}{d\,t} = -i_2\,R_c - \frac{i_4}{A} = -i_2\,R_c - \frac{q}{C} \qquad (7)$$

in which q is the instantaneous value of the charge on condenser C

But $\frac{d\,q}{d\,t} = i_3$ and $i_2 = i_3 + i_4$. So from (7)

$$\frac{1}{A}\frac{d\,i_4}{d\,t} = \frac{i_3}{C} \qquad (8)$$

And $$i_2 = \frac{C}{A}\frac{d\,i_4}{d\,t} + i_4 \qquad (9)$$

From 4, 5, 6, and 7,

$$R_p(i_1 + i_2) = -i_1 R_L - L\frac{d\,i_1}{d\,t} - \mu\,M\frac{d\,i_1}{d\,t} \qquad (10)$$

From which $$i_2 = -i_1\left(1 + \frac{R_L}{R_p}\right) - \left(\frac{L + \mu\,M}{R_p}\right)\frac{d\,i_1}{d\,t} \qquad (11)$$

From (9) and (11),

$$i_4 + \frac{C}{A}\frac{d\,i_4}{d\,t} = -i_1\left(1 + \frac{R_L}{R_p}\right) - \frac{L + \mu\,M}{R_p}\frac{d\,i_1}{d\,t} \qquad (12)$$

Using the second and third terms of (7), and substituting the value of i_2 from (9)

$$i_1 R_L + L\frac{d\,i_1}{d\,t} = i_4\left(R_c + \frac{1}{A}\right) + R_c\frac{C}{A}\frac{d\,i_4}{d\,t} \qquad (13)$$

Multiply (12) by R_c and use the value of $i_4\,R_c$ so obtained in equation (13)

$$\frac{i_4}{A} = i_1\left(R_L + R_c + \frac{R_c R_L}{R_p}\right) + \left[L + \frac{R_c}{R_p}(L + \mu\,M)\right]\frac{d\,i_1}{d\,t} \qquad (14)$$

Divide (12) by A, substitute for $\frac{i_4}{A}$ its value from (14)

$$\frac{C}{A^2}\frac{d\,i_4}{d\,t} = -i_1\left[R_L + \left(1 + \frac{R_L}{R_p}\right)\left(R_c + \frac{1}{A}\right)\right] - \frac{d\,i_1}{d\,t}\left[L + \left(R_c + \frac{1}{A}\right)\frac{L + \mu\,M}{R_p}\right] \qquad (15)$$

Derive (14), multiply by $\frac{C}{A}$, and combine with (15).

$$\frac{d^2\,i_1}{d\,t^2}C[L\,R_p + R_c(L + \mu\,M)] + \frac{d\,i_1}{d\,t}[C(R_L R_p + R_c R_p + R_L R_c) + L\,A\,R_p + (A\,R_c + 1)(L + \mu\,M)] + i_1[R_L\,R_p\,A + (R_p + R_L)(A\,R_c + 1)] = 0 \qquad (16)$$

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$$\frac{d^2 i_1}{dt^2}+\frac{d i_1}{dt}\frac{(C R_L+A L)(R_p+R_c)+C R_c R_p+L+\mu M(A R_c+1)}{C[L(R_p+R_c)+R_c \mu M]}+$$
$$i_1\frac{A R_L(R_p+R_c)+A R_p R_c+R_p+R_L}{C[L(R_p+R_c)+R_c \mu M]}=0 \qquad (17)$$

or
$$\frac{d i_1^2}{dt^2}+\frac{d i_1}{dt}\frac{a}{N}+i_1\frac{b}{N}=0 \qquad (18)$$

in which

$$a=(C R_L+A L)(R_p+R_c)+C R_c R_p+L+\mu M(A R_c+1)$$
$$b=A R_L(R_p+R_c)+A R_p R_c+R_p+R_L$$
$$N=C[L(R_p+R_c)+R_c \mu M]$$

In case the conductance of the condenser is zero and no resistance is used in the capacity branch ($A=0$, $R_c=0$).

$$a=C R_L R_p+L+\mu M$$
$$b=R_p+R_L$$
$$N=C L R_p$$

The solution of (18) will be a sinusoidal, damped, current if $\frac{a^2}{4N^2}<\frac{b}{N}$. If $\frac{a}{2N}$ is zero any oscillations started keep up indefinitely. For oscillations to build up $\frac{a}{2N}$ must be negative, which means that M must be negative, and in absolute value

$$M>\frac{1}{\mu(A R_c+1)}[L+C R_c R_p+(C R_L+A L)(R_p+R_c)] \qquad (19)$$

For the perfect condenser, with $A=0$ and $R_c=0$, oscillations will occur if M is negative and satisfies the condition for absolute magnitude,

$$M>\frac{1}{\mu}(L+C R_L R_p)$$

The frequency of the oscillations, if they do occur, is given from equation (18) by the relation

$$(2\pi f)^2=\omega^2=\frac{b}{N}-\frac{a^2}{4N^2} \qquad (20)$$

or for the perfect condenser,

$$f=\frac{1}{4\pi L}\sqrt{\frac{4L}{C}\left(1+\frac{R_L}{R_p}\right)-\left[R_L+\frac{1}{C R_p}(L+\mu M)\right]^2} \qquad (21)$$

The foregoing theory may be tested by opening the oscillatory circuit and putting it in a Wheatstone bridge, as shown in

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Figure 3. The voltage introduced for measuring. e_L. will be a sine wave, equal in magnitude to the drop in one arm of the bridge.

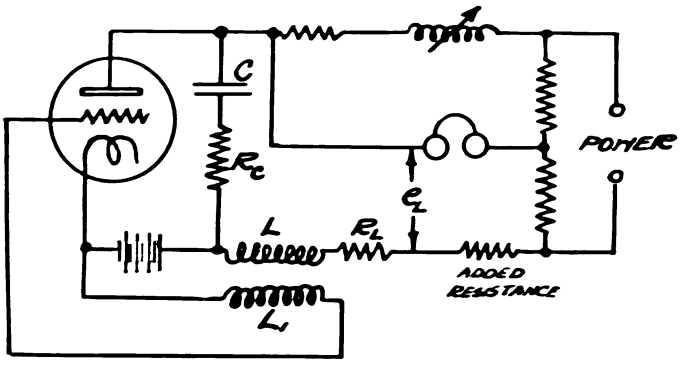

FIGURE 3

The derivation of the equation for this circuit is much the same as before, having, of course, the extra voltage, e_L. to take care of, however.

The equation, similar in form to (18), comes out

$$\frac{d^2 i_1}{dt^2}+\frac{d i_1}{dt}\frac{a}{N}+i_1\frac{b}{N}=\frac{1}{N}\left[\frac{d e_L}{dt}C(R_p+R_c)+e_L(R_pA+R_cA+1)\right] \quad (22)$$

Putting $e_L=E_L\varepsilon^{j\omega t}$, we know the current will be sinusoidal and representable by

$$i_1=F_1\varepsilon^{j\omega t}+jF_2\varepsilon^{j\omega t}$$

Getting the proper derivatives, substituting in equation (22) and equating real and imaginery terms, we find

$$F_1=E_L\frac{(b-N\omega^2)(R_pA+R_cA+1)+a\omega^2C(R_p+R_c)}{(b-N\omega^2)^2+(a\omega)^2} \quad (23)$$

$$F_2=E_L\frac{\omega C(R_p+R_c)(b-N\omega^2)-a\omega(R_pA+R_cA+1)}{(b-N\omega^2)^2+(a\omega)^2} \quad (24)$$

The impedance of the circuit Z_L is given by

$$Z_L=\frac{E_L}{I_1}=\frac{E_LF_1}{F_1^2+F_2^2}+j\frac{E_LF_2}{F_1^2+F_2^2}R=L'+j\omega L_L' \quad (25)$$

So

$$R_L'=\frac{(b-N\omega^2)(R_pA+R_cA+1)+a\omega^2C(R_p+R_c)}{(R_pA+R_cA+1)^2+\omega^2C^2(R_p+R_c)^2} \quad (26)$$

$$L_L'=\frac{a(R_pA+R_cA+1)-C(R_p+R_c)(b-N\omega^2)}{(R_pA+R_cA+1)^2+\omega^2C^2(R_p+R_c)^2} \quad (27)$$

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In case $R_c = 0$ (which was true in our experiments)

$$R_L' = \frac{R_L(R_p A+1)^2 + R_p(R_p A+1) + \omega^2 C R_p[L(A R_p+1) + C R_L R_p + \mu M - L(A R+1)]}{(R_p A+1)^2 + \omega^2 C^2 R_p^2}$$

$$= R_L + \frac{R_p(R_p A+1) + \omega^2 C R_p \mu M}{(R_p A+1)^2 + \omega^2 C^2 R_p^2} \qquad (28)$$

Putting in an equivalent series resistance, r, in place of A we have $\omega^2 C^2 r = A$ and the denominator of equation (28) becomes $R_p^2 \omega^2 C^2 (1+\omega^2 C^2 r) + 2 R_p r \omega^2 C^2 + 1$.

In the measurements we shall record we have, approximately, $R_p = 5{,}000$. $\omega = 8{,}000$, $C_{max.} = 10^{-6}$, $r_{max.} = 2$, $C_{min.} = 0.2 \times 10^{-6}$ So $\omega^2 C^2_{max.} r_{max.} \ll 1$, $2 r_{max.} \ll R_p$, $R_p^2 \omega^2 C^2_{min.} \gg 1$, so the denominator of equation (28) becomes approximately $R_p^2 \omega^2 C^2$.

Then
$$R_L' = R_L + r + \frac{\mu M}{C R_p} + \frac{1}{\omega^2 C^2 R_p} \qquad (29)$$

By a similar transformation we get, from equation (27),

$$L_L' = L - \frac{1}{\omega^2 C} + \left(\frac{r}{R_p} + \frac{1}{\omega^2 C^2 R_p^2}\right)\mu M$$

For resonance $L_L' = 0$ and we get from the above equation

$$\omega^2 = \frac{C R_p^2 - \mu M}{C^2 R_p (L R_p + r \mu M)} = \frac{1 - \frac{\mu M}{C R_p^2}}{L C + C \frac{r}{R_p} \mu M}$$

$$= \frac{1}{L C}\left[1 - \frac{\mu M}{R_p}\left(\frac{1}{R_p C} + \frac{r}{L}\right)\right] \qquad (30)$$

Above resonance, the circuit becomes capacitive and so a condenser must be used in the bridge to get a balance. This would open up the bridge circuit which has permitted the flow of continuous current, so we have to re-arrange the test circuit in accordance with Figure 4.

By derivations similar to those used for the circuit of Figure 3 we get, using R_c' to indicate the resistance as measured in the circuit of Figure 4.

$$R_c' = R_c + r + \frac{R_p R_L(R_p + R_L) + \omega^2 L R_p(L + \mu M)}{(R_p + R_L)^2 + \omega^2 (L + \mu M)^2} \qquad (31)$$

And as $R_L \ll R_p$ we may put

$$R_c' = R_c + r + \frac{R_p^2 R_L + \omega^2 L R_p(L + \mu M)}{R_p^2 + \omega^2 (L + \mu M)^2} \qquad (32)$$

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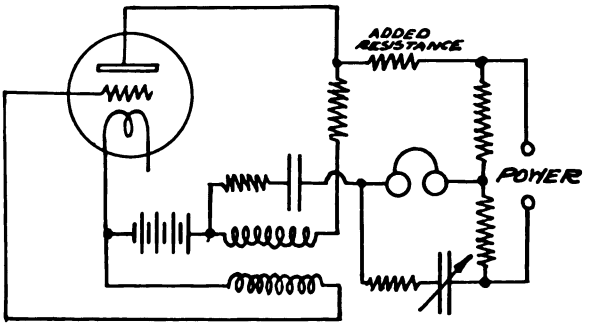

FIGURE 4

This resistance, it is to be noted, is independent of the amount of capacity used in the circuit.

Resonant frequency is given by

$$\omega^2 = \frac{1}{LC}\left(1+\frac{R_L}{R_p}\frac{\mu M}{L}+\frac{(L+\mu M)^2}{LCR_p^2}\right) \tag{33}$$

and the resistance as resonance is (if we put $R_c=0$)

$$R_c'(\text{resonance}) = R_L+r+\frac{L+\mu M}{CR_p} \tag{34}$$

For measurement in the inductive branch the resistance at resonance becomes, after legitimate approximation in equation (29)

$$R_L'(\text{resonance}) = R_L+r+\frac{L+\mu M}{CR_p} \tag{35}$$

which is just the same as it is in the capacitive branch. It is to be noticed however, from equations (30) and (33), that the resonant frequency is higher when measuring in the capacitive branch than when measuring in the inductive branch.

VARIATION OF RESISTANCE WITH FREQUENCY, FOR RESONANT CONDITION,

At resonant frequency

$$R' = R+\frac{L+\mu M}{CR_p} \text{ (by putting } R=R_L+r+R_c)$$

If we now make the ratio of inductances of the two coils (in Figures 3 and 4) constant, so that $L_1=m^2L$, if we keep the ratio of L to C constant so that $C=nL$, and as $M=K\sqrt{LL_1}$ we have, at resonance

$$R'(\text{resonance}) = R+\frac{L+\mu Kmn L}{nLR_p} = R+\frac{1+\mu Km}{nR_p} \tag{36}$$

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It thus seems that for a given tube the resistance of the oscillatory circuit will be independent of frequency, provided the ratio of L to C is kept constant, and a constant coupling is maintained between the coils, which are always selected with a definite ratio of inductances.

Thus suppose a certain combination gives -30 ohms resistance at $\omega = 16{,}000$. We change the size of wire used in the coils so as to get 4 times as many turns in the same size coil and keep the coils in the same relative position. The inductances will be increased 16 times and the coefficient of coupling will be the same as before. We increase the capacity 16 times. We drop the frequency so that $\omega = 1{,}000$ and the circuit will again have -30 ohms resistance.

Oscillatory Circuit Connected to Grid

The next circuit to investigate has the oscillatory circuit connected to the grid, as in Figure 5. Altho this circuit looks much like that previously used, the solution is somewhat different. We again assume that the grid current is sufficiently small to be neglected without appreciable error.

As preliminary equations we have

$$R_p i_p = e_p + \mu e_g$$

$$e_p = -M\frac{di}{dt} - L_1\frac{di_p}{dt}$$

$$e_g = -R_L i - L\frac{di}{dt} - M\frac{di_p}{dt}$$

$$= R_c i + \frac{i_2}{A} = R_c i + \frac{q}{C}$$

Now $\dfrac{dq}{dt} = i_1$, so $\dfrac{1}{A}\dfrac{di_2}{dt} = \dfrac{i_1}{C}$ and $i = i_1 + i_2$.

From these equations we get by combination and differentiation

$$\frac{di_p}{dt} = -\frac{1}{L_1}\left(e_p + M\frac{di}{dt}\right) = -\frac{1}{M}\left(e_g + L\frac{di}{dt} + R_L i\right) = \frac{1}{R_p}\left(\frac{de_p}{dt} + \mu\frac{de_g}{dt}\right) \tag{37}$$

Using the second and third members of (37)

$$e_p = \frac{L_1}{M}\left(e_g + L\frac{di}{dt} + R_L i\right) - M\frac{di}{dt} \tag{38}$$

Differentiating (38) and using this in the third and fourth members of (37)

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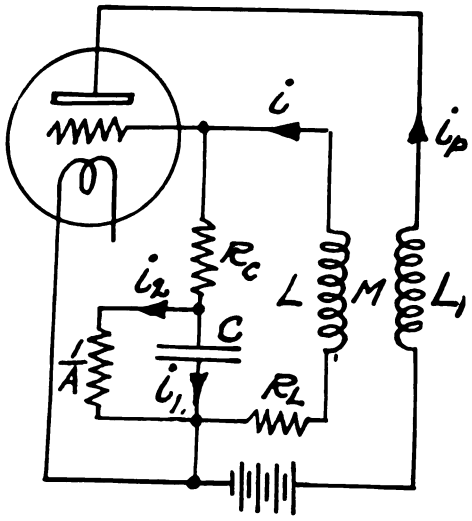

FIGURE 5

$$-\frac{1}{M}\left(e_g+L\frac{di}{dt}+R_L i\right)=\frac{1}{R_p}\left[\frac{L_1}{M}\left(\frac{de_g}{dt}+L\frac{d^2i}{dt^2}+R_L\frac{di}{dt}\right)-M\frac{d^2i}{dt^2}+\mu\frac{de_g}{dt}\right] \tag{39}$$

Combining terms

$$\frac{d^2i}{dt^2}(LL_1-M^2)+\frac{di}{dt}(L_1R_L+R_pL)+iR_pR_L=-R_pe_g-\frac{de_g}{dt}(L_1+\mu M) \tag{40}$$

Now $e_g=R_ci+\frac{i_2}{A}=i\left(R_c+\frac{1}{A}\right)-\frac{i_1}{A}=R_ci+\frac{q}{C}$ so

$$\frac{de_g}{dt}=R_c\frac{di}{dt}+\frac{i_1}{C}$$ hence

$$\frac{1}{C}e_g+\frac{1}{A}\frac{de_g}{dt}=\frac{i}{C}\left(R_c+\frac{1}{A}\right)+\frac{R_c}{A}\frac{di}{dt} \tag{41}$$

By solving for e_g and $\frac{de_g}{dt}$ in equation (40), we get

$$\frac{d^2i}{dt^2}A(LL_1-M^2)+\frac{di}{dt}(AL_1R_L+AR_pL+CR_cR_p)+i(AR_pR_L+AR_pR_c+R_p)=(CR_p-A(L_1+\mu M))\frac{de_g}{dt} \tag{42}$$

and

$$\frac{d^2i}{dt^2}(C(LL_1-M^2)+\frac{di}{dt}(CL_1R_L+CR_pL+CR_cL_1+CR_c\mu M)+i\{CR_pR_L+(AR+1)(L_1+\mu M)\}=-\{CR_p-A(L_1+\mu M)\}e_g \tag{43}$$

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Differentiating (43) and adding the resultant equation to (42), we finally get

$$\frac{d^3 i}{d t^3} C(L L_1 - M^2) + \frac{d^2 i}{d t^2}[A(L L_1 - M^2) + C(R_L L_1 + R_p L) + C R_c(L_1 + \mu M)] + \frac{d i}{d t}[A(L_1 R_L + L R_p) + (A R_c + 1)(L_1 + \mu M) + C R_p(R_c + R_L)] + i[A R_p(R_c + R_L) + R_p] = 0 \quad (44)$$

Or

$$\frac{d^3 i}{d t^3} + \frac{d^2 i}{d t^2}\frac{a}{N} + \frac{d i}{d t}\frac{b}{N} + i\frac{c}{N} = 0 \quad (45)$$

in which

$$a = A(L L_1 - M^2) + C(R_L L_1 + R_p L) + C R_c(L_1 + \mu M)$$
$$b = A(R_L L_1 + R_p L) + (A R_c + 1)(L_1 + \mu M) + C R_p(R_L + R_c)$$
$$c = A R_p(R_L + R_c) + R_p$$
$$N = C(L L_1 - M^2)$$

To test the theory of this circuit we imagine the circuit opened for measurement in a Wheatstone bridge as indicated in

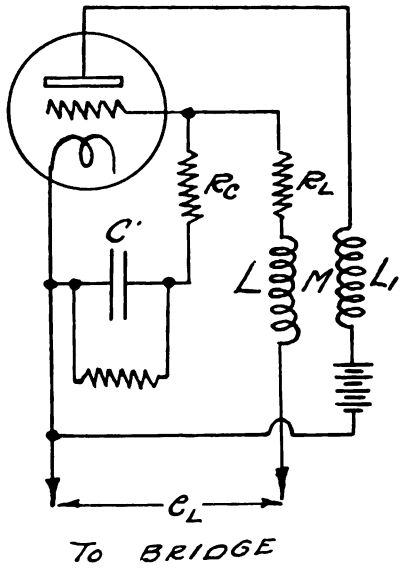

FIGURE 6

Figure 6. An analysis carried out to find the impedance to be measured in the bridge yields the equation

$$\frac{d^3 i}{d t^3} + \frac{d^2 i}{d t^2}\frac{a}{N} + \frac{d i}{d t}\frac{b}{N} + i\frac{c}{N} = \frac{d^2 e_L}{d t^2}\frac{C L_1}{N} + \frac{d e_L}{d t}\frac{C R_p + A L_1}{N} + e_L\frac{A R_p}{N}$$

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in which a, b, c, and N have the same meaning as before. Writing $e_L = E_L \epsilon^{j\omega t}$ as before and solving for the two components of current F_1 and F_2

$$F_1 = E_L \frac{(C R_p + A L_1)\omega^2(b - \omega^2 N) + (A R_p - \omega^2 C L_1)(c - \omega^2 a)}{(c - \omega^2 a) + \omega^2(b - \omega^2 N)^2}$$

$$F_2 = E_L \frac{\omega(C R_p + A L_1)(c - \omega^2 a) - (A R_p - \omega^2 C L_1)\omega(b - \omega^2 N)}{(c - \omega^2 a) + \omega^2(b - \omega^2 N)^2}$$

$$Z = \frac{E_L}{I} = \frac{E_L F_1}{F_1^2 + F_2^2} + j \frac{E_L F_2}{F_1^2 + F_2^2} = R_L' + j\omega L'$$

from which

$$R_L' = \frac{\omega^2(C R_p + A L_1)(b - \omega^2 N) + (A R_p - \omega^2 C L_1)(c - \omega^2 a)}{\omega^2(C R_p + A L_1)^2 + (A R_p - \omega^2 C L_1)^2} \quad (46)$$

$$L_L' = \frac{(A R_p - \omega^2 C L_1)(b - \omega^2 N) - (C R_p + A L_1)(c - \omega^2 a)}{\omega^2(C R_p + A L_1)^2 + (A R_p - \omega^2 C L_1)^2} \quad (47)$$

By substituting the values of a, b, c, and N, and making the same simplifying assumption as before ($A \ll \omega C$ and $R_c = 0$), we get

$$R_L' = R_L + r + \frac{R_p}{R_p^2 + \omega L_1^2}\left(\omega M^2 + \frac{\mu M}{C}\right) \quad (48)$$

and

$$\omega L_L' = \left(\omega L - \frac{1}{\omega C}\right) - \frac{\omega L_1}{R_p^2 + \omega L_1^2}\left(\omega M^2 + \frac{\mu M}{C}\right) \quad (49)$$

If the circuit is opened in the capacity branch, the resistance R_c' will be found independent of the value of the capacity, as it was for the first circuit analyzed (equation (32)).

The resonant frequency is given by the relation

$$\omega^4 L_1 C(L L_1 - M^2) - \omega^2(L_1^2 + \mu M L_1 - R_p^2 C L) - R_p^2 = 0 \quad (50)$$

If M is small, the resonant frequency is given approximately by the relation $\omega_o^2 = \frac{1}{L C}$.

Calling the resonant frequency (or more accurately $2\pi \times$ the frequency), ω, and putting $\frac{\omega_o^2 - \omega^2}{\omega^2} = x$ we get

$$x^2 + x\left(1 + \frac{\omega_o^2 L_1}{R_p^2}(L_1 + \mu M)\right) + \frac{\omega_o^2 L_1}{R_p^2}\left(\frac{M^2}{L} + \mu M\right) = 0 \quad (51)$$

Then using the general equations (48) and (49) (after putting $r = 0$), we find from equation (49) the value of

$$\frac{1}{R_p^2 + \omega L_1^2}\left(\omega M^2 + \frac{\mu M}{C}\right)$$

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and use this value in (48) and so get as the resistance of the circuit at resonance (for any value of M)

$$R_L'(\text{resonance}) = R_L + \frac{R_p}{\omega L_1}\left(\omega L - \frac{1}{\omega C}\right)$$

$$= R_L - R_p \frac{L}{L_1} x \qquad (52)$$

CAN M BE TOO LARGE TO PRODUCE OSCILLATIONS?

We determine the values of L_1 for which R_L' (resonance) is zero with the maximum possible value of M. We have

$$M = -K\sqrt{L\,L_1}$$

and using this in equations (51) and (52) we get (for R_L' (resonance) $= 0$).

$$L_1^2 - L_1^{3/2} K \mu L^{1/2} + L_1\left(\frac{R_p R_L}{\omega_o^2 L} + \frac{K^2 L R_p}{R_L}\right) - L_1^{1/2} \frac{R_p}{R_L} K \mu L^{3/2} + \frac{R_p^2}{\omega_o^2} = 0 \qquad (53)$$

In our tests we had $\mu = 6.75$, $L = 0.03$, $\omega_o^2 = 10^8$, $R_p = 5{,}000$, $R_L = 3$, and maximum available $K = 0.8$.

So $\quad L_1^2 - 0.935\, L_1^{3/2} + 32\, L_1 - 46.8\, L_1^{1/2} + 0.25 = 0$

which gives $\sqrt{L_1} = 1.425$ or $L_1 = 2.03$ henrys. So if we should put more than 2.03 henrys in the tickler coil, with $K = 0.8$, the set would not oscillate.

In a somewhat simpler derivation we can write (as approximately true) equation (53) in the form

$$K^2 \frac{R_p}{R_L} L\, L_1 = K \mu \frac{R_p}{R_L} L_1^{1/2} L^{3/2}$$

which gives

$$L_1 = \frac{\mu^2}{K^2} L$$

For $K = 0.8$ and $\mu = 6.75$ this gives the upper limit of L_1 as 70 times L, or if $L = 0.03$ the maximum value of L_1 which can be used is 2.1 henrys, which is practically the same as obtained by solution of the more complex equation.

In general $\omega L_1^2 \ll R_p^2$ so that equation (48) may be written

$$R_L' = R_L + \frac{1}{R_p}\left(\overline{\omega M}^2 + \frac{\mu M}{C}\right) \qquad (54)$$

As M is negative the effect of this term $\frac{\mu M}{C}$ is to give negative

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resistance and the effect of $\overline{\omega M^2}$ is to give positive resistance. Hence at some value of M the two terms will just neutralize and the effect of the triode on the resistance of the circuit is zero. The value of M at which this equality occurs will evidently increase as C decreases. It is given by

$$\omega^2 M = \frac{\mu}{C}, \quad \text{or} \quad M = \frac{\mu}{\omega^2 C} \tag{55}$$

Variation of Resonant Frequency as M Varies, M Being Large

If we consider $L_1 = 2$ henrys and $M = 0.2$ and the other constants same as given above, the resonant frequency will vary appreciably as M is varied. We have to solve equation (51) to get the ratio of the resonant frequency to that given by $\omega_o^2 = \frac{1}{LC}$ This effect will depend upon the frequency and was calculated for various values of M, and three frequencies; the results are

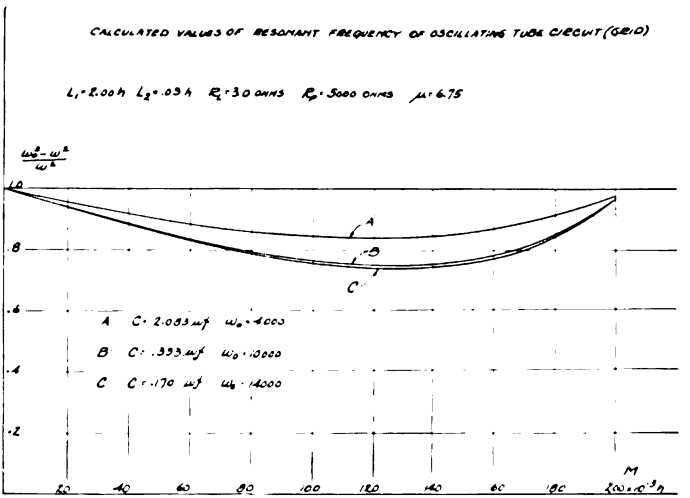

Figure 7

given in Figure 7. The value of the resistance of the oscillatory circuit, at the resonant frequency as fixed by the curves of Figure 7, is given in Figure 8. It is seen that this circuit, with the constants given, would oscillate until M approached the value of about 0.2 henry and with greater values of tickler coil (coupling being kept constant), the tube would refuse to oscillate.

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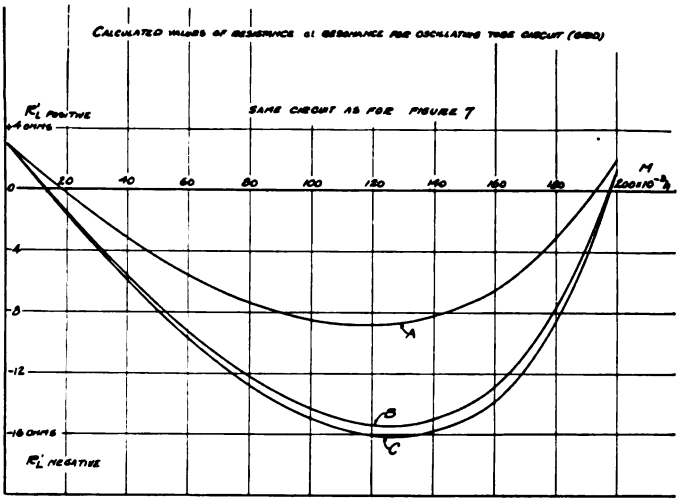

FIGURE 8

RESISTANCE AND REACTANCE OF THE CIRCUIT AS FREQUENCY IS VARIED

Using equations (48) and (49), the resistance and reactance of the oscillatory circuit were calculated, using the following conditions:

$R_L = 3$, $R_p = 5{,}000$, $\mu = 6.75$, $L = 0.03$, and for the values of C, 2.083 $\mu f.$, 0.333 $\mu f.$, and 0.170 $\mu f.$, which give resonance for $\omega = 4{,}000$, 10,000, and 24,000 radians, this covering the frequency range for which the circuit seemed suitable. Two values of L_1 were used.

Case *a*	$L_1 = 0.04$
Case *b*	$L_1 = 2.0$

The results, for resistance, are shown in Figures 9-11 and for reactance in Figures 12-14. The curves for reactance are nearly independent of the value of L_1, so only those for $L = 0.04$ henry are shown.

EXPERIMENTAL INVESTIGATION

The characteristics of the two triode circuits, the theory of which has been given in the previous paragraphs, were experimentally determined and the results are given herewith.

The circuit was arranged as in Figure 3, a Wheatstone bridge being used to get the resistance and reactance of the circuit. As long as the circuit had a sufficiently large capacity the impressed frequency (that of the power used on the bridge) was above the

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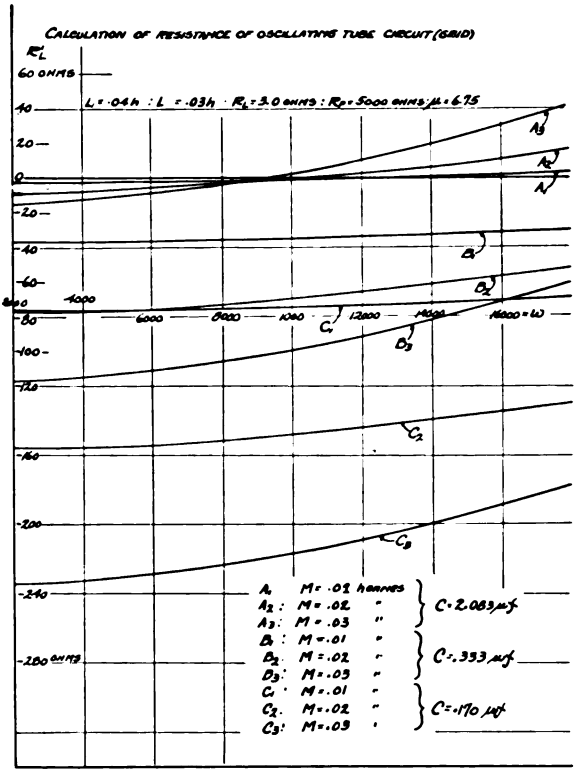

FIGURE 9

resonant frequency of the circuit, its reactance was inductive, and a variable inductance was used in the bridge to get the reactance balance. With capacity sufficiently small to make the impressed frequency lower than the natural frequency of the circuit it was necessary to use a condenser to balance the bridge. As the plate current of the tube would, in this case, have to flow thru the telephones (the capacity arm of the bridge being open) and as this would ruin the telephones and make the plate voltage excessively low, the oscillatory circuit was opened in the capacity branch, thus permitting the plate current to flow to the plate without going thru the bridge at all. This is indicated in Figure 4.

A vacuum tube oscillator, giving a frequency of about 1,200 cycles was used as a source of power for the bridge. Measurements were made on the circuits for about fifteen values of M between zero and its maximum possible value, this representing about 80 percent coupling. Various condensers, the capacity and series resistance of which had been accurately determined,

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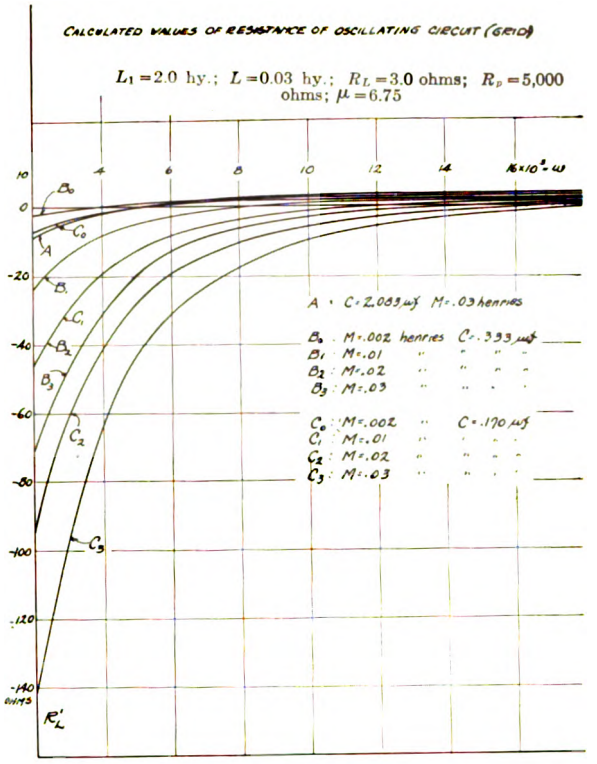
CALCULATED VALUES OF RESISTANCE OF OSCILLATING CIRCUIT (GRID)
L_1 = 2.0 hy.; L = 0.03 hy.; R_L = 3.0 ohms; R_p = 5,000 ohms; μ = 6.75
ω x 10⁻³
A : C = 2.083 μf M = .03 henries
B_0 : M = .002 henries C = .333 μf
B_1 : M = .01 " " " "
B_2 : M = .02 " " " "
B_3 : M = .03 " " " "
C_0 : M = .002 " C = .170 μf
C_1 : M = .01 " " " "
C_2 : M = .02 " " " "
C_3 : M = .03 " " " "
OHMS
R'_L

FIGURE 10

were used in the oscillatory circuit; their values ranged from seven microfarads to one-tenth microfarad. The value of the equivalent series resistance of the condensers involves both R_c and A of the theory, as no extra resistance was used in the capacity branch.

The experimental results have a precision of better than 1 percent in most cases; in the only cases where a greater discrepancy than this occurred it was later found that the tube had ionized and had thereby changed the plate circuit resistance. When first starting the tests the tube had a resistance (alternating current) of 5,115 ohms, but during the progress of the work, it ionized a few times, and its final value of resistance was only 4,300 ohms.

After performing a series of measurements, with fixed value of C and varying M, the oscillatory circuit, where it had been opened for bridge measurement, was closed, and the value of M was increased until oscillations started, to see how closely the

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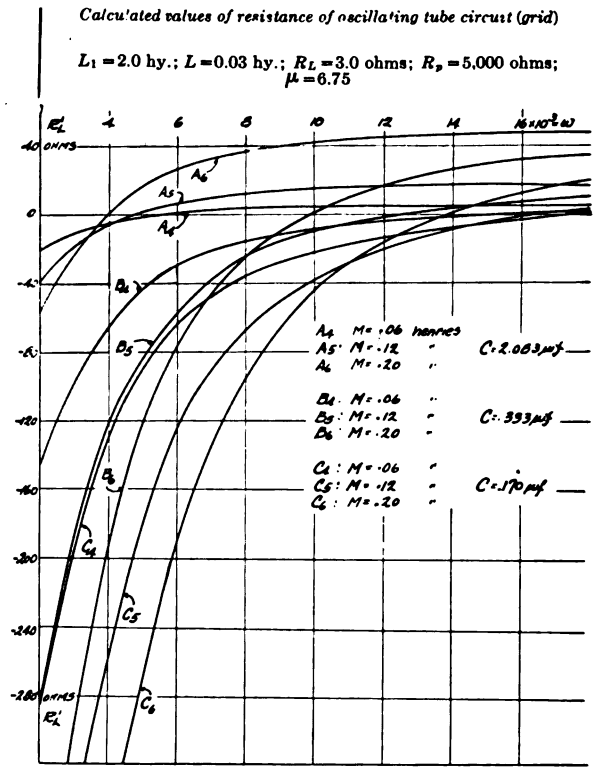
Calculated values of resistance of oscillating tube circuit (grid)
L_1 = 2.0 hy.; L = 0.03 hy.; R_L = 3.0 ohms; R_p = 5,000 ohms; μ = 6.75
OHMS
A_6
A_5
A_4
B_4
B_5
B_6
C_4
C_5
C_6
A_4 M = .06 henries
A_5 M = .12 "
A_6 M = .20 "
C = 2.083 μf
B_4 M = .06 "
B_5 M = .12 "
B_6 M = .20 "
C = .393 μf
C_4 M = .06 "
C_5 M = .12 "
C_6 M = .20 "
C = .170 μf

FIGURE 11

condition of zero resistance corresponded to the condition for starting oscillations. As free oscillations must occur at a frequency fixed by the circuit constants, it is necessary that the circuit show negative resistance at the circuit's natural frequency before oscillations will start.

Thus altho the bridge measurement might show negative resistance, oscillations might not start when the circuit was closed because at the *natural* frequency the resistance may be positive. But if, in the bridge, the circuit showed zero reactance and negative resistance, it would always start to oscillate as soon as the circuit was closed.

Thus when the impressed frequency was the same as the natural frequency and the coupling was sufficiently weak that its effect on the natural frequency was negligible, the total resistance of the oscillatory circuit as measured in the bridge might look like that in Figure 15; if, with the oscillatory circuit closed, the coupling was increased until M was equal to 5.2 millihenrys, the circuit would at once begin to oscillate.

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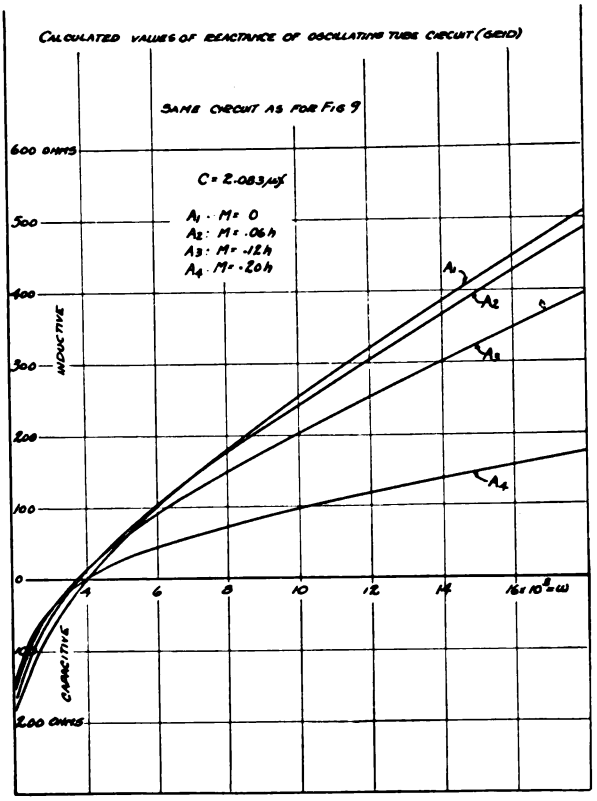

FIGURE 12

The results of the bridge measurements are shown in Figures 16-19. In Figure 16 are shown the curves of the resistance of the oscillatory circuit when this is in the plate circuit of the triode and the bridge is in the inductive branch of the circuit. The resistance for this case is given theoretically by equation (29); this is seen to predict a linear relation between resistance and mutual inductance and the experimental results show this to be so. The equation predicts that, for circuit constants as used in this test, the smaller the capacity the tighter must the coupling be made to reduce the circuit resistance to zero, and the curves substantiate the theory.

Curve 8 of Figure 16 shows the results when just enough capacity was used in the oscillatory circuit to make this resonant for the impressed frequency, so that the bridge showed zero reactance. With such a capacity the tube circuit will start to oscillate by itself, at this frequency, if the value of mutual in-

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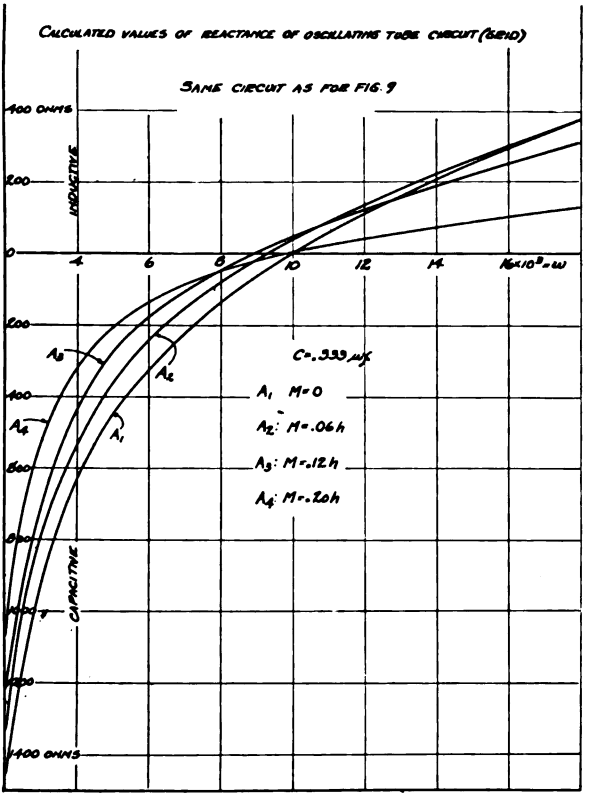

FIGURE 13

ductance is increased to 5.5 millihenrys or more. Having in mind the slight discrepancy between theory and experiment, due to the approximations made in simplifying the theory, experiment checks exactly with theory. As nearly as could be measured, oscillations started at once when the value of M demanded by equation (29), to make R_L' equal to zero, was reached.

The values of resistance used in plotting the curves of Figure 16 have had the equivalent series resistance of the condenser taken out, as this was entirely different for the different condensers used. The values plotted on the curve sheet, therefore, are $R_L'-r$, where R_L' is the circuit resistance measured in the bridge and r is the equivalent series resistance of the condenser used.

For the highest value of coupling used ($M=27$ millihenrys) the values of $R_L'-r$ have been calculated from equation (29), for the various values of C used. These values, together with the

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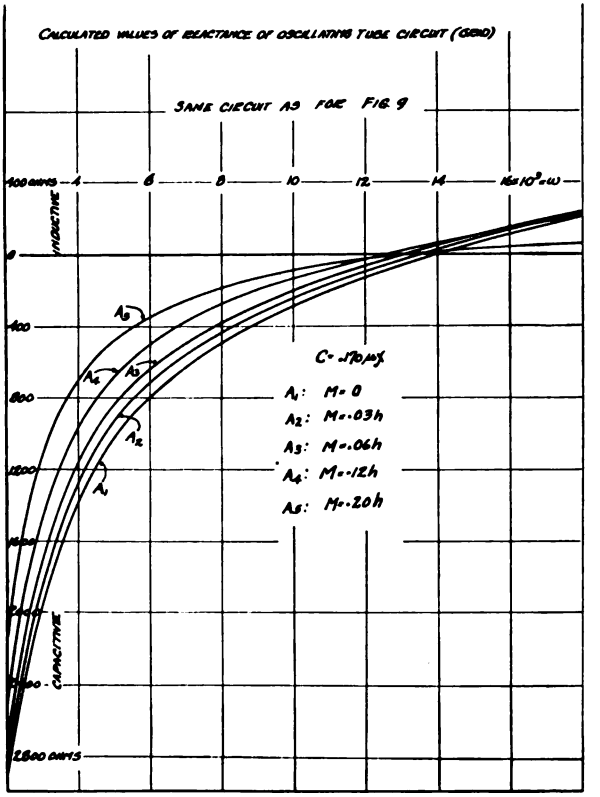
CALCULATED VALUES OF REACTANCE OF OSCILLATING TUBE CIRCUIT (SEND)
SAME CIRCUIT AS FOR FIG. 9
INDUCTIVE
CAPACITIVE
C = 470 μμf
A1: M = 0
A2: M = .03h
A3: M = .06h
A4: M = .12h
A5: M = .20h

FIGUEE 14

experimentally determined values, have been tabulated on the curve sheet. The agreement can be seen to be remarkably close, especially when it is remembered that 200 ohms of resistance was permanently left connected in the oscillatory circuit to prevent oscillations while measurements were being made. The agree-

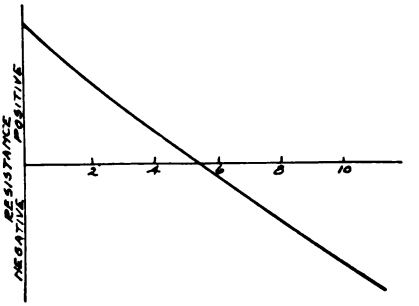

FIGURE 15

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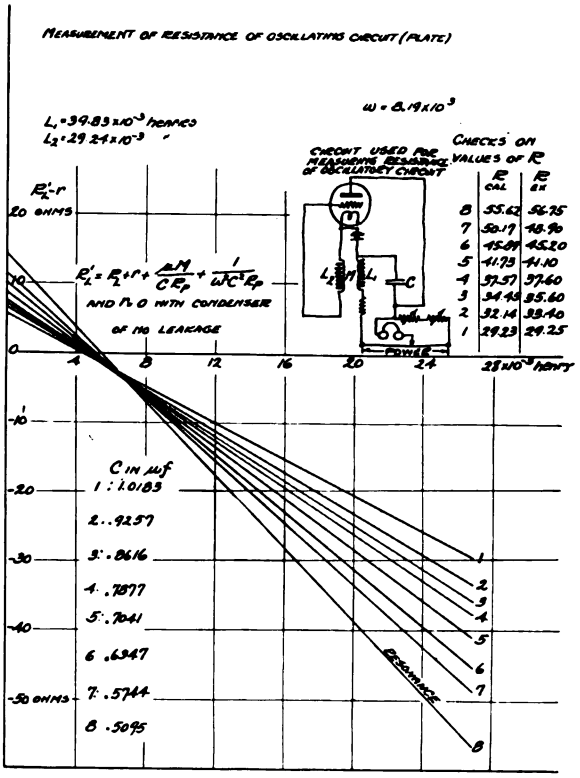

FIGURE 16

ment of theory with the experimentally determined values is generally closer than ½ percent.

When the capacity used was greater than that required for resonance it was necessary to introduce the bridge in the capacity branch of the oscillatory circuit as noted before; the theoretically derived resistance for this case is given by equation (32). This equation predicts that the resistance should be independent of the capacity used in the oscillatory circuit and such is seen to be the case for the experimental curves given in Figure 17. The equivalent series resistance of the condenser has again been subtracted from the measured value of circuit resistance. Here also the divergency of the experimental values from those predicted by theory is very small in nearly all cases, being much less than 1 percent.

When the oscillatory circuit is connected to the grid, theory shows the resistance should vary with M as given in equation (48). In Figures (18) and (19) are shown curves of the measured value

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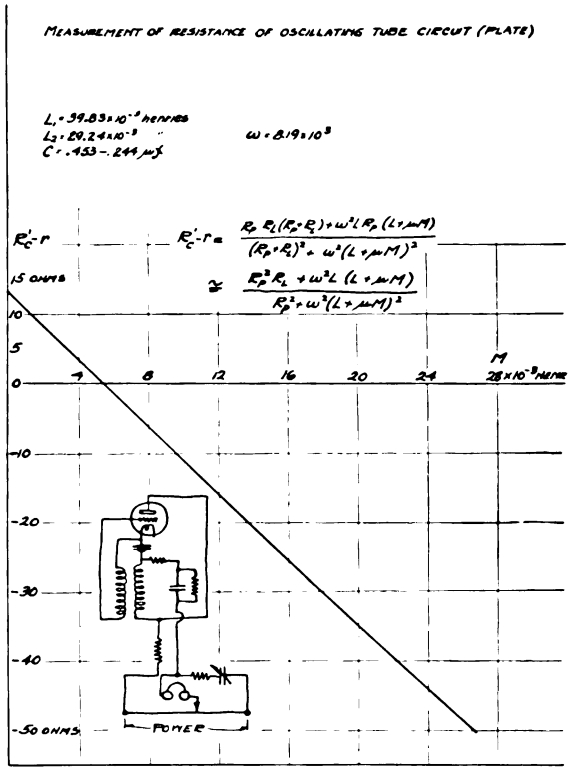

FIGURE 17

of the resistance of the circuit after $(R_L + r)$ has been taken out, these curves then give the experimentally determined values of the third term of the right-hand side of equation (48). It is here seen that the effect of coupling is much different than was the case when the oscillatory circuit was connected to the plate. The coupling introduces negative resistance into the oscillatory circuit for low values of M (this being negative in polarity) and positive resistance for high values of M. The amount of negative resistance developed in the circuit varies with the capacity used much more than was the former case. In Figure 18 the resistance is shown for the larger condensers used to bring out clearly the peculiar form of this curve and in Figure 19 are shown curves for the complete range of condensers used in the experiments; for the smallest capacity used, 0.0903 microfarad, the negative resistance generated with tightest coupling, exceeded 400 ohms. Other circuits with properly adjusted constants give thousands of ohms of negative resistance.

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It is, of course, to be always remembered that the negative resistance cannot actually exist as the sole resistance of the circuit; such a circuit would in general be unstable. These values of negative resistance are the values given by the interaction of the tube circuits and generally exist only if the positive resistance of the circuit is greater than this negative value, that is, the resistance of any stable circuit, as a whole, must be positive.

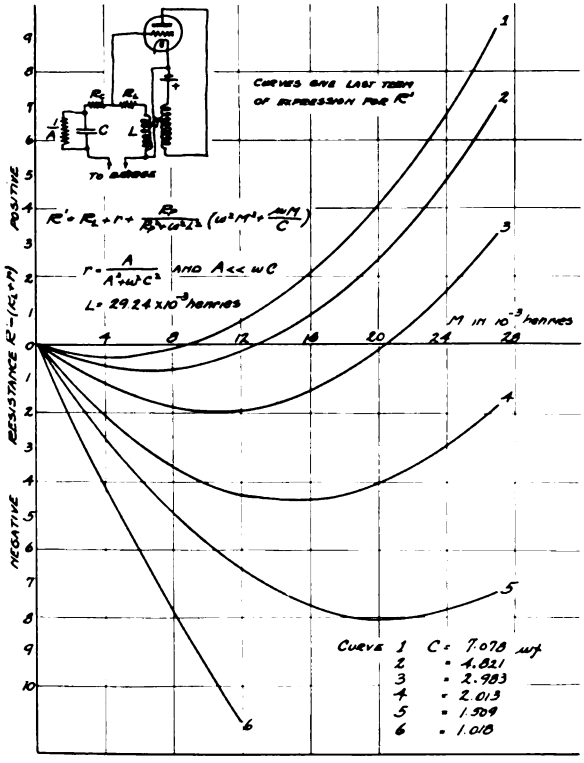

FIGURE 18

The experimental determinations given in this test for the oscillating condition were generally good to four significant figures, a precision easily obtainable with the alternating current bridge; the departure of these values from the theoretical values was always such as could have been predicted, being due to approximations in the theory or to a slow change in circuit constants as the tests progressed. Also R_L and r were determined only for one frequency and the measuring frequency differed

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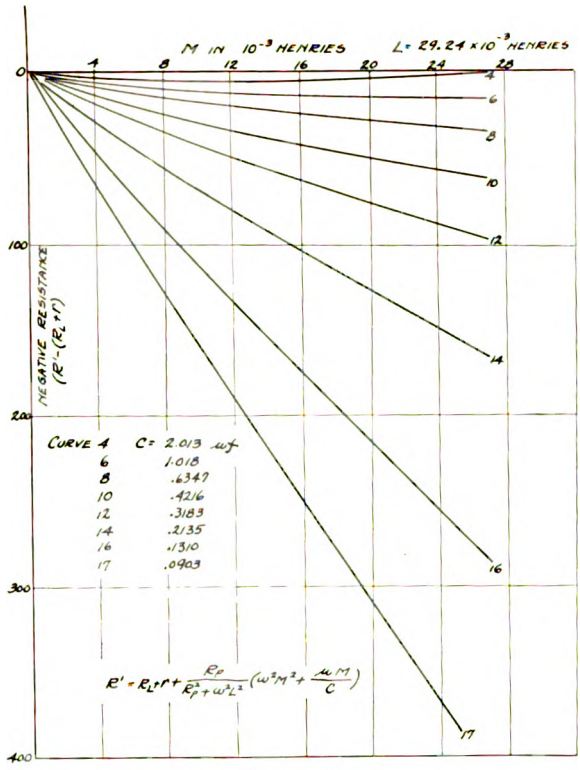
M IN 10^{-3} HENRIES
$L = 29.24 \times 10^{-3}$ HENRIES
NEGATIVE RESISTANCE $(R' - (R_L + r))$
CURVE 4 C = 2.013 mf
6 1.018
8 .6347
10 .4216
12 .3183
14 .2135
16 .1310
17 .0903
$R' = R_L + r + \frac{R_p}{R_p^2 + \omega^2 L^2}\left(\omega^2 M^2 + \frac{\mu M}{C}\right)$

FIGURE 19

from this. Greater precision is possible, but it is felt that the results given fully substantiate the theory of these two circuits.

SUMMARY: Two typical triode circuits, used for obtaining regeneration by mutual induction, are examined analytically, with the idea of testing the results of the analysis by experiment.

As the experiments were to be performed with a one thousand cycle current (because of the greater accuracy of measurement at this frequency) it was permissible to neglect the condensive coupling due to the tube elements, and furthermore, as the conditions for the *starting* of oscillations were to be determined, the current taken by the grid could be neglected as it is only one or two microamperes in the non-oscillatory state.

Measurements made on the circuits, by using the alternating current bridge, confirm the theoretically derived relations to a remarkable degree of precision, showing that triode circuits are susceptible of accurate analysis.

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DISCUSSION

Lewis M. Hull (by letter): The painstaking analysis and ingenious experimental methods set forth in this paper deserve praise. But it appears to me that the writers have directed their efforts away from those problems of triode operation which are of importance in modern radio practice. Their analyses are concerned solely with the first-order effects in two typical circuits: the tuned-grid and tuned-plate circuit, with inductive feed-back coupling. By first-order analysis I mean, in general, consideration of quasi-plane sections of the triode characteristic surface, which leads to the usual linear circuital equations with *constant* coefficients. The results of investigations of the two circuits discussed in this paper are applicable mainly to two general processes in tuned triode circuits: first, regenerative amplification of externally impressed oscillations, and second, self-generation of sustained oscillations. If we turn back to the literature on this subject, we find that the elementary results obtainable by first-order analysis alone have been derived again and again. With regard to regenerative amplification we find the classical papers of Vallauri, seven years old ("Elettrotecnica," **4**, 335, 350, 1917), of Ballantine (this journal, **7**, 2, page 155, 1919), and, more recently, the thoro analysis of Chaffee (this journal **12**, 3, 299, 1924). In fact, the framework of the typical first-order analysis of regenerative circuits has become essentially text-book material for modern engineering courses, in a class with the alternating-current transformer and the induction motor. As far back as 1918, moreover, Hazeltine carried the first-order analysis to its useful limit in the problem of generation of sustained oscillations by a triode (this journal, **6**, 2, 63, 1918). All of this work reaches the usual conclusions on negative resistance, conditions for starting oscillations, variations of national frequency with circuit constants, etc. To be sure, no previous investigator has measured those particular first-order effects described in the present paper. But the point I wish to emphasize is that it is unfortunate that much talent and diligence should still be applied, in the present day, to ferreting out the details of various circuit combinations on the basis of the old linear equations, when the real problems, which combine theoretical interest with immediate practical importance, are concerned almost wholly with the departures of the current-voltage surfaces of the triode from simple planes.

For instance, the radio technician is not so vitally interested in the properties of the negative resistance which a feedback

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coupling introduces into a tuned circuit *provided* the assumption of linear characteristics is fulfilled: Armstrong discussed this negative resistance pretty thoroly without any reference to the calculus at all. The engineer wants to know why the ultimate current obtainable in a tuned receiving circuit without local oscillation is not independent of the original resistance of that circuit, as the linear negative-resistance equations predict. The answer to such a question can be found only by assuming the resistance of some element of the circuit to be a function of the *amplitude* of *oscillation*, which is the condition that actually obtains in most practical circuits. I have investigated this question mathematically, and have obtained an approximate solution of the equations for a variable grid-filament resistance (appropriate in the case of a regenerative detector tube) and for a variable plate-filament resistance, R_p By considering variations with the amplitude in either of these two triode coefficients alone, it is possible to explain the experimental phenomenon of a progressive decrease in the generative amplification as the original or basic loss resistance in the tuned circuit is increased.

In the case of the triode generator it has been shown repeatedly, not only that the slope of the characteristic surface at the operating point (expressed in terms of amplification "constant" and plate-resistance) has little significance in determining the amplitude of oscillation, but that stable oscillations are actually impossible on plane sections of this surface. I wish to point out that the literature contains several *complete* discussions of the whole problem of oscillation with a triode having the necessary curved characteristics. Many years ago Barkhausen supplied a volume of analysis on the oscillating arc, much of which is directly applicable to the triode problem ("Das Problem der Schwingungserzeugungen," Pries, Leipzig, 1907). Van der Pol ("Radio Review" I, 701, 1920), gave an excellent discussion of the non-linear problem, using Lorentz's method of perturbations, but without experimental verification. Appleton and Van der Pol ("Phil. Mag.," 42, 201, 1921), carried the analysis further and obtained experimental verification of their results on triode oscillations over a wide curved range of the voltage-current characteristics. Two years ago I published in these PROCEEDINGS (**10**, 5, 373, 1922), a complete mathematical theory of sustained oscillations in a triode generating circuit which are of such wide amplitude as to extend far into the saturation region of the characteristic surface, and thus apply to conditions in which the triode is most efficient as a power converter. I consider that my method is

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perhaps superior to Van der Pol's, insofar as it permits either exactness or approximation, as the occasion demands, and allows a prediction of the alternating-current power obtainable from a tube, and of the circuit constants necessary for maximum power output or maximum efficiency, all of which have been checked repeatedly in experimental practice. The extra cooling of a power-tube filament, due to the latent heat of evaporation of electron current, was considered, with the resulting errors which this effect introduces into the "static characteristics" of a power triode. Accordingly, a bridge method was developed for measuring at low frequencies the "generation constants" of a triode, from which its power output in any mono-periodic generating circuit are calculated. The methods described in my paper are of most value in connection with triodes of small power capacity; I have never applied them in practice to tubes larger than the conventional "five-watt" type. But the outstanding point is that the elementary non-linear viewpoint has long since been passed, in application to practical operation of triode circuits. In view of the quantity of theoretical and concordant experimental work which has already been published on the curved characteristics of the triode, the real "gaps" which remain to be filled in do not appear to be touched by the present paper. Undoubtedly many other investigators in this country have worked out methods for handling the curvature of triode characteristics in the circuit equations; they would perform a further scientific service by publishing their results in these PROCEEDINGS.

J. H. Morecroft (by letter): It seems as tho Dr. Hull had failed to read the first paragraph of our paper, otherwise most of his reiteration is difficult to understand. The excellent work of Armstrong, Hazeltine, Ballantine, himself, and other investigators, as published in the PROCEEDINGS, does not give experimental proof of the theory (insofar as I can see), and, unimportant as it may seem to a research man of Dr. Hull's type, the experimentally obtained curves, substantiating the theory, are often of more value to the general worker and student than the theory itself.

Not all papers can be original in the sense that they deal with a subject never touched before—not all of us can be Faradays or Hertz's. In 1919, I published in the "American Institute of Electrical Engineers Proceedings," a paper on the triode as a power converter, giving operating curves exactly like those given by Dr. Hull in his 1922 paper in these PROCEEDINGS, and many

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other curves perhaps more important than those Dr. Hull gave three years later. But that did not make his paper of no value! Five years ago, in my radio text, I discussed in an elementary way the effects of the variation of the plate circuit resistance of a triode, but that doesn't make valueless any contribution Dr. Hull may add, due to research activities he is carrying out at present.

The mathematically inclined research worker will do well to remember that not all people think as he does; some have to progress by routes which involve more detailed effort and repetition than he thinks worth while.

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EXPERIMENTAL DETERMINATION OF THE FUNDAMENTAL DYNAMIC CHARACTERISTICS OF A TRIODE*

By

Eijiro Takagishi

(Radio Engineer, Electrotechnical Laboratory, Ministry of Communication, Tokyo, Japan)

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I. General Considerations

Because of the recent development in vacuum tube technique, not only the field of radio communication but also several other branches of science have made such rapid progress that one could hardly have foreseen these wonderful changes even a few years ago. Nevertheless, progress has been rather slow in the experimental determination of the real functioning of the triode.

The writer believes that more efficient and correct utilization of the triode can be attained only thru exact knowledge of its characteristics. Up to the present, the characteristics of the triode in its static condition have been much investigated; while on the other hand, when dealing with its performance as a generator or an amplifier of oscillating current or potential, the dynamic characteristics become important in order fully to understand the operation of the system.

As is well known, the current flowing in a triode in actual operation is not free from distortion, and large harmonic com-

* Received by the Editor, February 25, 1924.

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ponents are present in addition to the fundamental. When a triode is used in an electric circuit which can be actually tuned to a resonant frequency, the relation between the various electrical quantities is ordinarily important only insofar as it relates to their fundamental frequency.

With this idea in mind the writer carried out some experiments for measuring the fundamental components of electrical currents in triode circuits under conditions similar to those found in practice. The results are described below, but it is very regrettable that the writer is unable to give them quantitatively, because all of the records of the investigations were destroyed during the recent conflagration in Tokyo.

II. Methods of Measuring Fundamental Frequency Components in Distorted Electric Currents

Three methods may be considered for measuring the fundamental component in a distorted electric current. These are:

(a) Tuned circuit method
(b) Wattmeter method
(c) Vibration galvanometer method.

In the first method, there is used a tuned circuit consisting of inductance and capacity, the natural frequency of which circuit is equal to that of the current to be measured. This method is convenient for the measurement of currents of higher frequencies but is not so practically suitable for that of lower frequencies.

The second method utilizes the following characteristics of a wattmeter. When sinusoidal currents of the same frequency are made to flow thru both of the two coils of a wattmeter, it always indicates their product, and even when one of them is replaced by a distorted current, the wattmeter indication remains unchanged so far as the fundamental of the distorted current does not change both in amplitude and phase. The wattmeter can be calibrated by using a sinusoidal current obtained from an ordinary power source and it can then be readily used for the measurement of the fundamental component of any distorted current.

This method is ingenious in theory and also simple and convenient for experiment, but it should be borne in mind that even a slight distortion in the auxiliary electric current causes a large error, especially when the current to be measured contains a large proportion of harmonics. For example, only five percent of the third harmonic in the auxiliary current and three hundred

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percent in that to be measured give rise to an error of fifteen percent in wattmeter reading.

The last method depends upon the principle of alternating current potentiometer, using as an indicator a vibration galvanometer, which is resonant to a fundamental frequency.

When a sinusoidal current i_s flows thru a conductor A (Figure 1), and the current i_d which is to be measured flows thru the other conductor B, and a vibration galvanometer is bridged across the two resistances R and r (which are inserted in each conductor), the galvanometer will cease to vibrate at a definite value of r/R. Then the fundamental component of i_d is found to be $i_s \frac{R}{r}$. By the use of a galvanometer which does not respond to any other harmonics, the zero point will remain unchanged as long as the amplitude of the fundamental in i_d is kept constant. Even in the case when the ammeter reading of i_s is used directly, instead of its fundamental component, the percentage error is actually small. For example, if i_s contains five percent of harmonic components, the error as a whole has a value of only 0.12 percent.

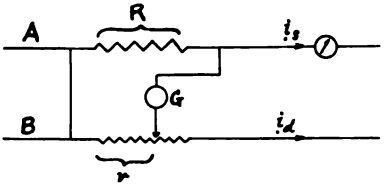

FIGURE 1

III. Measuring Arrangement for the Determination of the Fundamental Dynamic Characteristics of a Triode

Suppose that a triode V (Figure 2) behaves as a generator of radio frequency current, impedances Z_g and Z_a being placed in the grid and anode circuits.

If the voltage drops in Z_g and Z_a are e_g and e_a, we may substi-

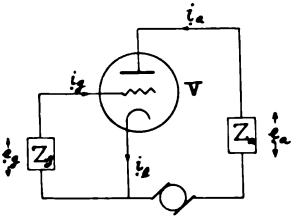

FIGURE 2

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tute for each of the impedances a generator of an emf., which is equal to that across the respective impedances with respect to both amplitude and phase. The practical experimental arrangement, which is based on this idea, is shown in Figure 3, the following abbreviations being used:

V—triode, the characteristics of which are to be studied,

B_a—battery or direct current generator in anode circuit,

B_g—battery in grid circuit,

B_k—battery in cathode circuit,

T_a—transformer to supply anode circuit with e_a,

T_g—transformer to supply grid circuit with e_g,

T_l—transformer in the local circuit,

F_g—phase transformer in grid circuit,

F_l—phase transformer in local circuit,

r_a, r_g, r_l—resistances in anode, grid, and local circuits, respectively,

r_p—potentiometer,

A_a, A_g, A_k, A_l—ammeters in anode, grid, cathode, and local circuits, respectively,

V_g, V_a—voltmeters in grid and anode circuits,

G—vibration galvanometer,

S—change-over switch,

t—ground connection.

The phase transformer is such that single phase voltage in any phase may be drawn from the rotor, three-phase voltage being applied in the stator and the rotor being turned. The electromotive forces e_a and e_g are applied to the anode and grid circuits by means of the transformers T_a and T_g. There then flows in each circuit a current of complicated wave form, which becomes more distorted as the applied voltage grows. The fundamental alternating components in the distorted currents are denoted by i_a and i_g. When i_a and i_g are made to flow thru r_a and r_g, the voltage drops are proportional to the currents flowing. The currents are then measured by potentiometer method, the drops being balanced to the sinusoidal voltage drop in r_p. The phase transformer F_l supplies, thru a transformer T_l, a current of pure sine form flowing in a local circuit which includes a potentiometer resistance r_p.

We first adjust the local current to a certain amount (for example, one ampere). The vibration galvanometer G is bridged across r_a and a part of r_p. When the galvanometer just ceases to

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vibrate, the fundamental voltage drops in r_a and the included part of r_p are equal to each other, that is, $\dot{i}_a$ is found by the following equation

$$\dot{i}_a = i_l\left(\frac{r_p}{r_a}\right), \tag{1}$$

where $\dot{i}_a$ signifies here the fundamental of the distorted alternating current and i the sinusoidal current in the local circuit.

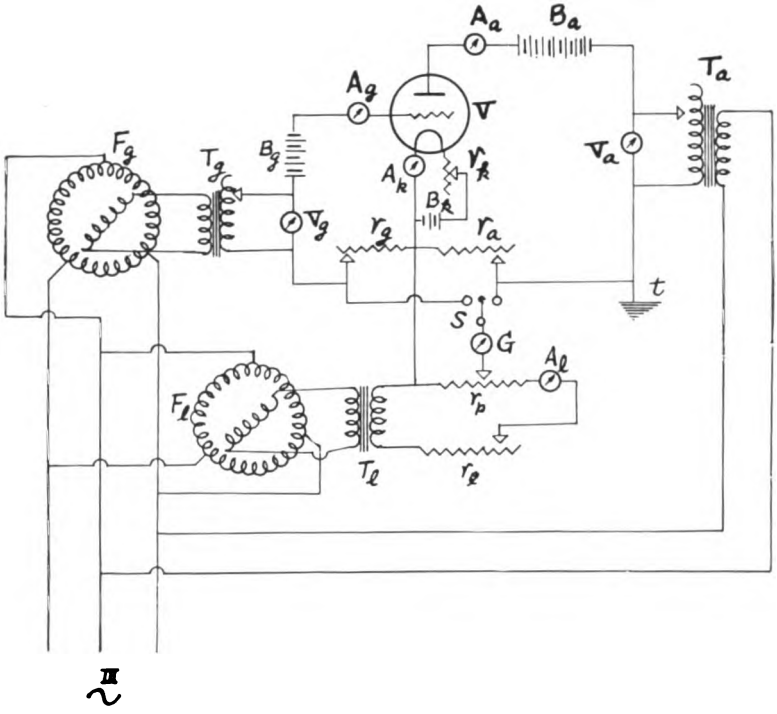

FIGURE 3

In cases where both amplitude and phase of $\dot{e}_a$ and $\dot{e}_g$ are varied to various extents, the current $\dot{i}_a$ can be measured easily and correctly by adjusting F_g, F_l, and so on. The distorted current $\dot{i}_g$ in the grid circuit can also be measured similarly. The writer has made some experiments with the case where $\dot{e}_a$ and $\dot{e}_g$ are in opposite phase, the arrangement being shown in Figure 4, which was chosen because he did not possess a phase transformer F_g, as required in the arrangement of Figure 3. A transformer T was used, which has two sections, S_a and S_g in the secondary, the former being for the anode and the latter for the grid circuit. L_{g1} and L_{g2} are for adjustment of the phase of e_g; L_{g1} serves to

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produce a lag and L_{g_2} a lead relative to the terminal voltage at S_g. Resistance R_g serves for fine adjustment of the amplitude of e_g.

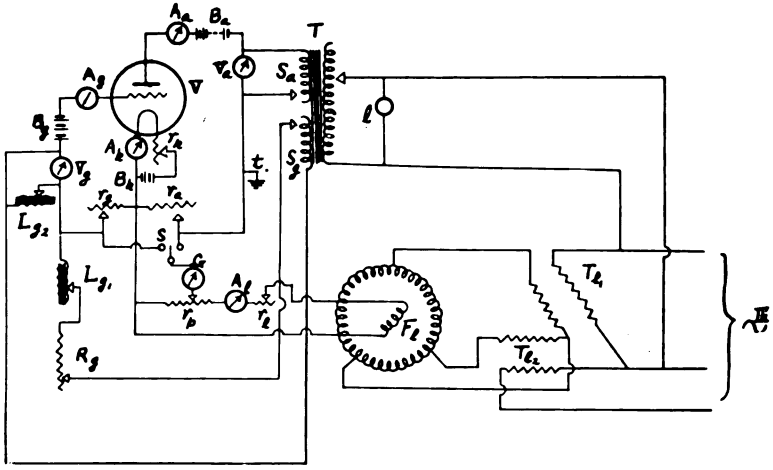

FIGURE 4

As the phase transformer F_l, the writer made use of a direct current shunt motor of one-half horsepower, the whole winding of its armature being divided into three symmetrical sections in order to render it available for three-phase supply, and two new windings, each having twenty turns, being placed on the existing field coils. 100 volts of three-phase supply were too high for safe direct application to the armature, and two transformers were used, by which 100 volts were stepped down to about 20 volts. Thus the induced 8 volts in the secondary was nearly cotnstan for all positions of the rotor. This emf. produced an electric current of less than one ampere in the local circuit. The instruments and arrangements used will be explained below in more detail.

Referring to Figure 4:

T—primary 100 volts, 6 steps, secondary voltages finely adjustable, secondary max. 1,000 volts, S_g max. 250 volts,

l—100-watt incandescent lamp, used for suppression of self-oscillation of the transformer T associated with the triode V,

T_{l_1}, T_{l_2}—single-phase transformer, primary 100 volts, secondary 20 volts,

V_a—alternating current voltmeter, max. 1,200 volts.

V_g—alternating voltmeter, hot-wire type, 100 volts, max., about 600 ohms resistance,

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A_a—direct current milli-ammeter, 500 milli-ampers max.,
A_g—direct current milli-ammeter, 150 milli-amperes max.,
A_k—direct current ammeter, 5 amperes, max.,
A_f—alternating current ammeter, 1 ampere max.,
r_p—5 ohms, about 50 cm. (19.4 inches) long, straight, B. and S. number 32,
r_f—20 ohms, 2 amperes,
r_a—22 ohm—44 ohm—220 ohm—and 440 ohm—taps producing 2.2 volts of potential drop by 100 m.a., 50 m.a., 10 m.a., and 5 m.a. of anode alternating current, respectively.
r_g—similar construction to r_a,
R_g—8 ohms, 2 amperes,
L_{g_1}—small iron core inductance coil for fine adjustment of phase,
L_{g_2}—large iron core inductance coil for fine adjustment of phase,
B_a—dry battery, max. 1,500 volts,
V—triode, manufactured by the Tokyo Electric Company, Ltd., 250 watts, cathode current 3 amperes,
B_k—accumulators, 20 volts,
G—vibration galvanometer, re-constructed from a 800-ohm telegraph relay; instead of the tongue, a vibrator of thin iron plate, was used, this tongue being about 0.08 mm. (0.03 inch) thick, adjusted to vibrate vigorously at a resonant frequency of 50 cycles, and operative down to 0.1 volt or 0.05 milli-ampere.

The current in the local circuit was maintained constant at 0.44 ampere in order that fifty centimeters (19.7 inches) in length of the potentiometer resistance wire may have a voltage drop of just 2.2 volts. If 22 ohms of resistance are used for r_a, then i_a is measured by

$$\left.\begin{aligned} i_a &= 100 \times \frac{(\text{length of resistance wire } r_p \text{ in cm.})}{50 \text{ cm.}} \\ \text{with 44 ohms for } r_a \quad i_a &= 50 \times \frac{(\text{length of resistance wire } r \text{ in cm.})}{50 \text{ cm.}} \\ \text{with 220 ohms} \quad {}_{\cdot a} &= 10 \times \frac{(\text{length of resistance wire } r \text{ in cm.})}{50 \text{ cm.}} \\ \text{and with 440 ohms} \quad i_a &= 5 \times \frac{(\text{length of resistance wire } r \text{ in cm.})}{50 \text{ c.m.}} \end{aligned}\right\} \quad (2)$$

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For the measurement of larger currents, the current $\dot{i}_l$ was increased to 0.66 ampere.

The process of measurement will be explained for the above-mentioned arrangement. We first keep the triode V in its normal condition, for example, with 750 volts of direct anode potential, 2.8 amperes of cathode current, and we then apply voltages from the transformers to the anode and grid so as to make $\dot{e}_a$ and $\dot{e}_g$ opposite in phase to each other. $\dot{e}_g$ and $\dot{e}_a$ cannot always be kept in exactly opposite phase, since the phase varies according to the amount of load on the transformers. To take care of this condition, L_{g1} and L_{g2} serve to adjust the phase precisely. We then throw the switch S to the left, slide the contactor of the potentiometer, and at the same time turn the rotor of the phase transformer F_l until the galvanometer ceases to vibrate, at the same time keeping the alternating current $\dot{i}_l$ in the local circuit at a constant value. We then throw the switch S to the right and observe whether the null point of the galvanometer be found on the potentiometer or not. Generally this is not the case, for $\dot{i}_a$ and $\dot{i}_g$ are not exactly in opposite phase. In that case, it is necessary to adjust the phase of $\dot{e}_g$ by L_{g1}, L_g, R_g, and so on, and find anew the desired condition. For example, if $\dot{e}_g$ lags a little behind $\dot{e}_a$, $\dot{i}_a$ also lags behind it. It is then necessary to make $\dot{e}_g$ lead but a little in order that $\dot{i}_a$ be in phase with $\dot{i}_l$; in that condition, however, the phase of $\dot{e}_a$ does not coincide with that of $\dot{e}_g$, but leads it by a small amount. Thus if $\dot{e}_g$ leads quite a little, the desired condition will readily be found, where all the voltages and currents $\dot{e}_a$, $\dot{e}_g$, $\dot{i}_l$, and so on, are in the same phase. Under these conditions, we must throw the switch to the left and then to the right, and the null point will be found in both cases, and thus $\dot{i}_a$ and $\dot{i}_g$ may be measured directly on the scale of the potentiometer.

Typical results of the measurement are shown in Figures 5, 6, 7, 8, and 9.

IV. Dynamic Characteristics of a Triode

Figure 5 shows the relations between effective values of the fundamental components of anode current (A), grid current (G), and their total (E) against applied grid voltage at a definite ratio of $\dot{e}_g$ and $\dot{e}_a$, which latter are in exactly opposite phase. It is interesting to note that curve A has a peak, past which the curve falls rapidly, while curve G begins to rise suddenly at that point and reaches a high value. This indicates that, in such conditions, the grid absorbs the larger part of the electrons emitted from the incandescent cathode.

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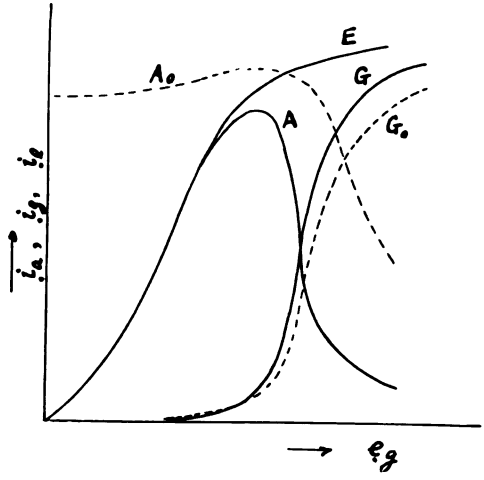

FIGURE 5

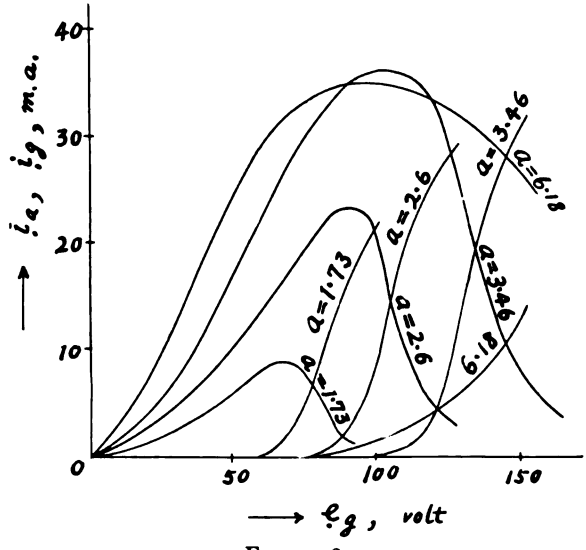

FIGURE 6

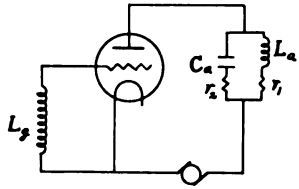

FIGURE 7

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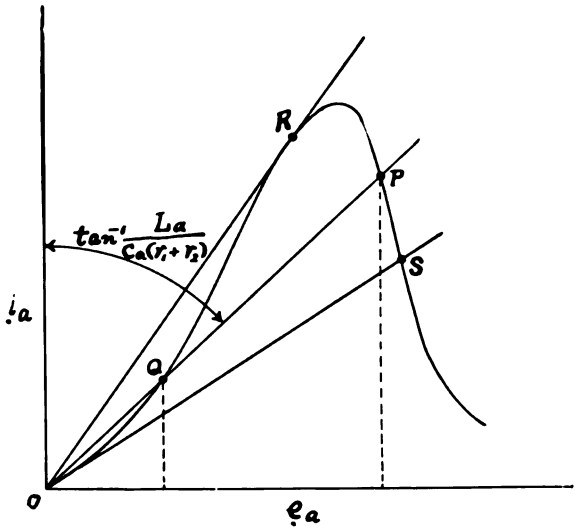

FIGURE 8

As to curve E, however, it never drops. In the figure, it is seen that grid current varies widely, depending upon the alternating voltage applied to the anode, and in the following equation for grid current a factor ν cannot be neglected,

$$i_g = g\ (e_g + \nu e_a) \tag{3}$$

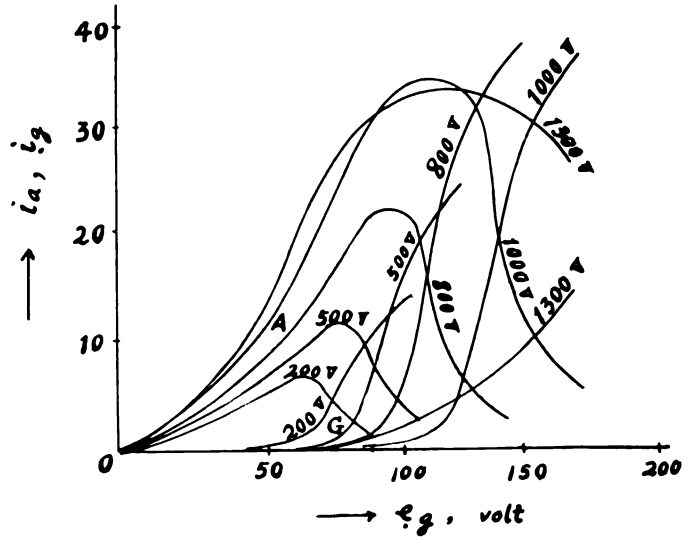

FIGURE 9

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The presence of grid current and the factor ν are both important in studying the operation of a triode. Their exact relations are not yet worked out.

Dotted lines A and G in the figure show the corresponding values of direct current in the anode and grid circuit. Curves A and G in the figure change in form for different values of e_g/e_a, as shown in Figure 6. The values of the ratio e_g/e_a multiplied by μ is denoted by a, where μ is so-called voltage amplification factor. In the present case, μ was measured and found to have a nearly constant value of 26, and a was selected between 1.73 and 6.18.

The characteristic curves explained above will be of great service for the predetermination of the stable oscillating condition of the triode. As a simple example, let a system oscillate with the circuit arrangement as shown in Figure 7, using a triode, the characteristic of which is known to have the form shown in Figure 8 for a certain value of a. In this case, the expression $\mu \frac{M}{L_a}$ can be used approximately for a, M being the mutual inductance between L_a and L_g.

Strictly speaking, the wave length at which the system is maintained in oscillation is not exactly equal to the value calculable when considering inductance and capacity only, but in the present case, for the sake of simplicity, the only capacity besides the condenser capacity which was considered in the determination of the wave length was that of the lead wires, which eliminated an appreciable error. Let C_a signify the resultant total capacity. Then the oscillating circuit consisting of L_a, C_a, r_1, and r_2 offers an equivalent resistance equal to $L_a/C_a\,(r_1+r_2)$, at the wave length calculated from L_a and C_a. We draw the straight line $O\ Q\ P$ (Figure 8), which has a slope equal to $tan^{-1}L_a/C_a(r_1+r_2)$, and intersects the dynamic curve at two points P and Q. One of these two points, P corresponds to the condition found in practice, where the values of e_a and i_a, and consequently that of the oscillation current are definite. If C_a is made larger and larger, the straight line $O\ Q\ P$ rises more and more steeply, and at a point R becomes tangent externally to the curve, and there oscillation dies away. To start oscillation again, it is necessary to give the straight line a greater slope, in other words, to make C_a smaller. Then the system begins to oscillate at a point S, $O\ S$ being tangent internally to the curve. In this case e_a and consequently e_g have rather larger values than those in the former case, but so far as the amplitude of the oscil-

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lation current in the resonant circuit is concerned, it is not proportional to e_a, but is rather smaller, because then the equivalent resistance has a larger value.

This phenomenon is already known as" Ziehen"[1] in Germany, and has been investigated in that country altho the investigators do not use dynamic characteristics as explained above, with the exception of Dr. Hans Georg Möller,[2] who, however, did not show the falling characteristics for the anode.

From the characteristics of a triode and its associated circuits, it will be fully understood that this phenomenon accompanies changes in wave length and that it takes place only when the characteristic has an inflection point on the left-hand side. The values of L_a, M, L_a/C_a (r_1+r_2), and so on, being measured and calculated, the writer predetermined, with C_a, $M_,$, and so on, as variables, the values of the oscillation currents according to the methods explained above and observed good agreement. As the direct voltage E_a on the anode is varied between 200 volts and 1,300 volts, the corresponding alternating anode and grid currents vary according to the curves in Figure 9, by which means of which modulation in the anode circuit can be predetermined. If the heating current thru the cathode is increased, those curves grow higher, as might be expected. If E_g were varied, similar curves might be expected. Unfortunately the great fire in Tokyo intervened before these measurements were started. It is regrettable that the results are, for that reason, given only qualitatively (against the writer's inclination).

The writer, however, believes that he will be able to publish the further details of the experiments later on. The above-mentioned instance is only for the special case where e_a and e_g are in opposite phase, while in other cases, matters are not so simple, but become more interesting and important.

V. Voltage Amplification Factor or Constant

The anode current is usually expressed by the following equation

$$i_a = h\,(e_a + \mu e_g) \tag{4}$$

where h is the variable conductance of a triode and μ is a factor called "voltage amplification factor," which is supposed to be constant with respect to a triode. In the equation it is seen that i_a is always zero when e_a plus e_g is equal to zero, and following

[1] "Ziehen" may be translated as "pulling" (intermittently) or "tugging."

[2] Dr. Hans Georg Möller, "Die Elektronenröhren und ihre Technischen Anwendung."

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this idea the writer measured this factor with the arrangement already mentioned, the results being shown in Figure 10. The dotted lines show the relation between e_a and e_g under the condition that i_a is zero, that is to say, the slope of any straight line $O\,P$ defines the value of μ in that condition.. The lines are curved as shown, in other words $\bar{\mu}$ is not constant.

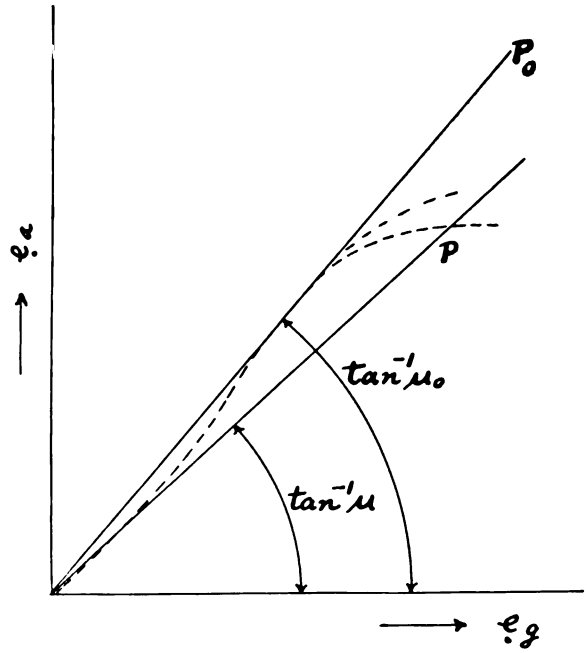

FIGURE 10

The cause of this is merely the existence of grid current, which is proved by the following experiments. The writer assumed the correctness of the equation

$$i_l = g_a\,(e_a + \mu_a\, e_g) \qquad (5)$$

where i_l is the fundamental component of current emitted from the cathode, and g_a and μ are the factors just defined. The measurement reveals the fact that the relation between e_a and e_g under the condition that i_l is zero is shown by a straight line $O\,P_0$, which remains unaffected by great changes in the values of the variables. Thus the newly established constant of voltage amplification is really constant for a given triode. The writer desires to express his obligation to Mr. E. Yokoyama, under whose direction he carried out the present experiments, and also to

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express his deepest indebtedness to Mr. S. Kawazoe for his valuable assistance thruout the whole series of experiments.

Synopsis

1. In studying the action of a triode, the static characteristics have been much considered, but little attention has been paid to the fundamental dynamic characteristics, which show the relation of much distorted currents and potential drops in a triode with reference to their fundamental components only.

2. A method of measuring the fundamental component in distorted electric current was adopted to determine the above mentioned characteristics of a triode. The experiment was confined to a special case, where grid and anode voltages are kept exactly in opposite phase, because all the apparatus and the arrangements had been destroyed by the recent conflagration in Tokyo before further experiments were executed. The circuitual arrangements of the measurements are fully described. Some of the fundamental dynamic characteristic curves of a triode thus determined are shown and explained.

3. It has been found by the experiment that the anode current-grid voltage characteristic curve has a falling region in a working condition and that the current which flows thru the grid circuit generally has a large value.

4. Some characteristics of a triode in stable oscillation were predetermined by the help of the characteristic curves and were in a good coincidence with those directly measured, the triode being operated as an oscillator. The phenomenon of "Ziehen" was briefly mentioned.

5. It was observed in the experiments that a voltage amplification factor or constant as at present defined is not of constant value, but is subjected to variation with applied voltages. A new constant has been established, which relates to the emission current from the incandescent cathode of a triode instead of the anode current.

6. It is much regretted by him that the writer is unable to give the results of the present researches quantitatively because of the recent conflagration at Tokyo, but he expects to go ahead with the experiment in the near future.

SUMMARY: By picking out the fundamental component only from the distorted current in a triode while functioning, the writer has found first new forms of dynamic characteristics for a triode, which are utilizable for solving various problems not hitherto satisfactorily solved (for example, modulation in radio telephony, the phenomenon of "Ziehen," or "pulling," and so on), and secondly that the grid current may reach a large value and that the factor or constant ν merits attention. Some experiments have been carried out relative to the voltage amplification factor for which a new expression has been found. The work is being continued.

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A HIGH EFFICIENCY VACUUM TUBE OSCILLATING CIRCUIT*

By

D. C. Prince and F. B. Vogdes

(Research Laboratory, General Electric Company, Schenectady, New York)

Introduction

In preceding issues of the Proceedings of The Institute of Radio Engineers, methods of calculating the performance of vacuum tube oscillators have been developed.[1] Except in Chapter VII of the cited papers, sinusoidal voltages only were considered. Chapter VII develops the method of calculating the increase in efficiency and output when the circuit is made responsive to more than the fundamental frequency.

A circuit called the "push-pull" circuit has been developed in which two tubes are operated 180° out of phase. Several circuits of this sort have also been suggested in which the idea that each tube should carry current half of the time and that the current form should be square in each tube (the summation current being held constant by an inductance in the supply lead), has been added to the two-tube idea.

The writer has spent considerable time reducing this form of circuit to a design basis. In doing this it has been necessary to make several changes in the original circuit. The present paper develops this study to a point where intermediate frequency circuits of very high efficiency and large specific output can be designed with reasonable certainty that they will work as desired.

Square Wave Oscillator

With the usual type of sine-wave oscillating circuit, current is drawn thru the tube for a very small part of the cycle when high efficiency is sought. The reason for this is that, since the plate voltage is varying sinusoidally, it will remain at a minimum for a very short time. During the remainder of the cycle the

*Received by the Editor, December 31, 1923.

[1] D. C. Prince, "Vacuum Tubes as Power Oscillators," Proceedings of The Institute of Radio Engineers, volume 11, numbers 3, 4, 5; June, August, and October, 1923.

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drop is high so that any current passed is accompanied by large loss. However, since the purpose of this discussion is to develop the design of a circuit of high output we will deliberately increase the angle during which current flows and then study ways of reducing the losses. Since a given thermionic valve has a definite amount of emission, the largest input will be obtained by taking the full emission while the current flows. For reasons which will appear later, the circuit will be made symmetrical so that each of two tubes may draw current of constant value during one-half of the cycle. Such a circuit is shown in Figure 1. This circuit of Figure 1 is not usefully operative, so no grid control is shown. Figure 2 shows the plate drop e_p, plate current i_p, and

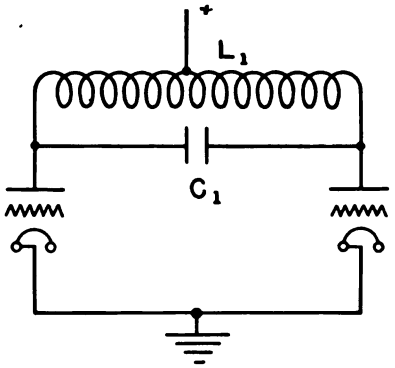

FIGURE 1

loss w_p for one of the two tubes. It is apparent that the losses at points "a" are much higher than at points "b," because the voltage drop across the tube is high at those points. At points "b" the drop need be only sufficient to overcome the space charge of the tube.

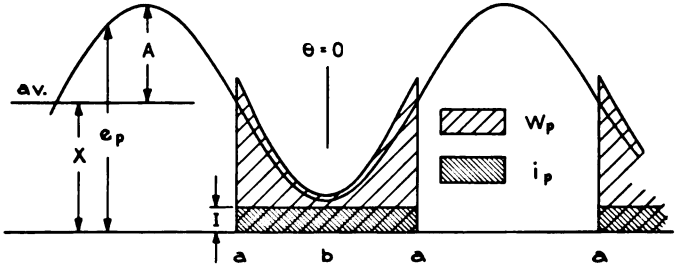

FIGURE 2

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The power input to the circuit is the product of current I and volts X, both of which are constant. The plate voltage is

$$e_p = X - A \cos \theta$$

Instantaneous loss is $W = e_p I = I (X - A \cos \theta)$

$$\text{Efficiency} = \frac{I X - \frac{1}{\pi} I \int_{-\frac{\pi}{2}}^{+\frac{\pi}{2}} [X - A \cos \theta] \, d\theta}{I X}$$

$$= \frac{A}{\pi X} \int_{-\frac{\pi}{2}}^{+\frac{\pi}{2}} \cos \theta \, d\theta$$

$$= \frac{2A}{\pi X}$$

Thus the efficiency is 0.637 times the ratio of counter emf· maximum to impressed direct voltage. A is always less than X.

Since we are trying to improve the efficiency without changing the current wave, we must proceed by changing the voltage wave. From Figure 2 it appears that the loss could be reduced to a minimum if the plate voltage e_p were kept at the minimum value necessary to overcome space charge for the entire period during which current flows. This would mean a voltage wave such as shown in Figure 3.

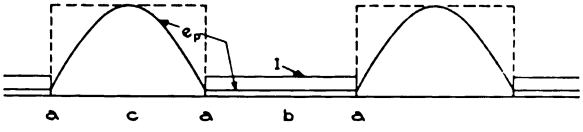

FIGURE 3

It is quite immaterial what the plate voltage is for the period "a c a," since no current is flowing. The important thing is to have e_p a minimum for the period "a b a" during which current is flowing. Whether the values of e_p should follow the full or dotted line for the period "a c a" is determined by the method used to control the potential in the period "a b a."

By the method described in Chapter VII, of the previously cited papers, a symmetrical square voltage wave could be obtained by employing a considerable number of trap circuits tuned to various odd harmonics. The circuit would, however, be extremely complicated and difficult to handle.

The circuit of Figure 1 may be added to as shown in Figure 4, by the addition of the choke coil L_3. If L_3 be

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made very large it will maintain the current at a constant value regardless of any cyclic voltage change. If the grid excitation is such as to pass this constant current for one-half cycle thru each tube with low plate potential, the remaining drop must be absorbed by the inductance L_3. Thus, with proper grid excitation, the voltage e_p could vary in accordance with the full line of Figure 3. Figure 5 shows the various voltage rela-

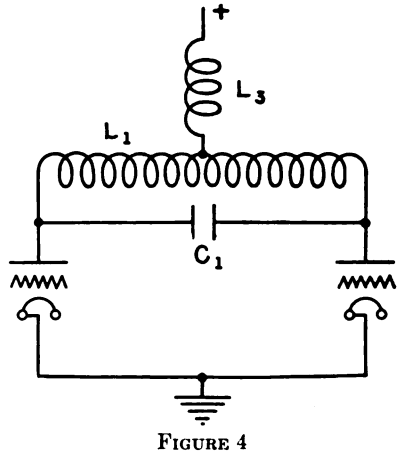

FIGURE 4

tionships, e_{p_1} and e_{p_2} are the plate potentials of the two tubes while e_{L_3} is the average potential at the mid-point of L_1. X is the average value of e_{L_3}, since there can be no constant difference of potential across an inductance such as L_3. The difference between e_{p_1} and e_{p_2} is the potential difference across L_1 and C_1 and is sinusoidal, since L_1 and C_1 constitute a resonant circuit. With this arrangement the energy put into the circuit at times "a" is stored in L_3 and does not appear as tube loss. At points "c" and "b" it is passed on to the oscillating circuit with no loss of efficiency.

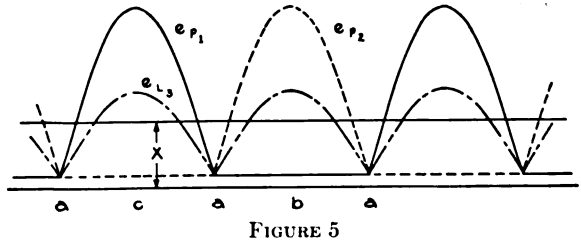

FIGURE 5

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The necessary condition to secure this highly efficient voltage relationship is that the tube control must not limit the current while current is flowing. The valve must be either wide open or entirely closed. Examination of the tube characteristic, Figure 6, shows that plate current and grid current are roughly proportional. A control system which will produce square blocks of grid current will then presumably produce square blocks of plate current at constant plate volts. Square blocks of grid

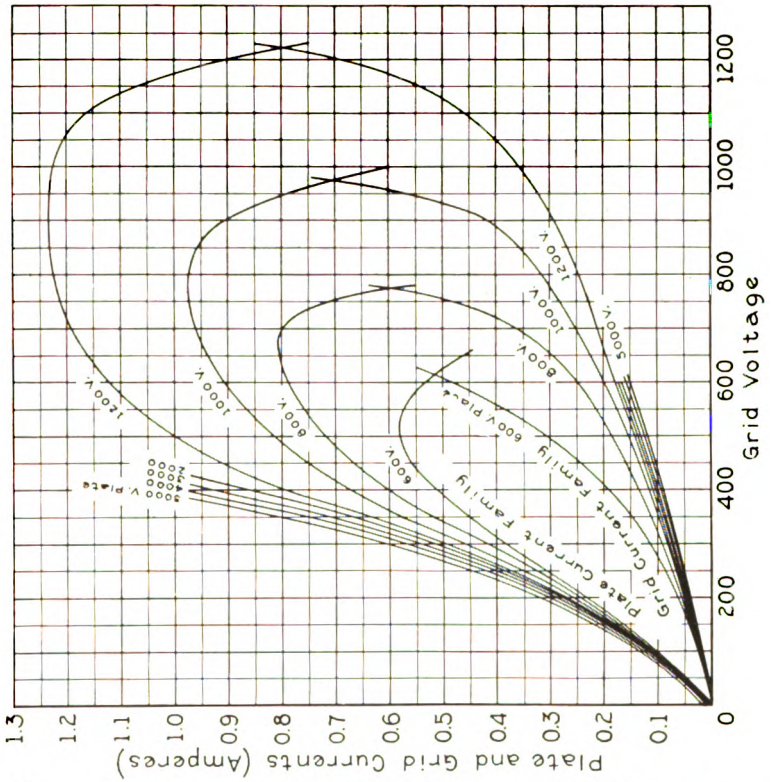

FIGURE 6—Characteristics of a UV-206 Radiotron with High Emission

current can be obtained by the arrangement shown in Figure 7. Energy is supplied to the grid by coupling between L_1 and L_2. The total grid current is kept constant by inductance L_4.

This grid current naturally flows from that grid which is most positive at any instant, and since the total is constant, each grid receives current in square blocks as required.

If L_3 and L_4 are not large enough to keep the total current constant in grid and plate circuits, there will be cyclic "ripple"

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variations in each of these circuits as shown in Figure 8. As in Figure 5, X is the impressed potential while e_{L_3} is the potential of the mid point of L_1 and represents the counter emf. of the oscillator. At points "a" the counter-electromotive force is zero.

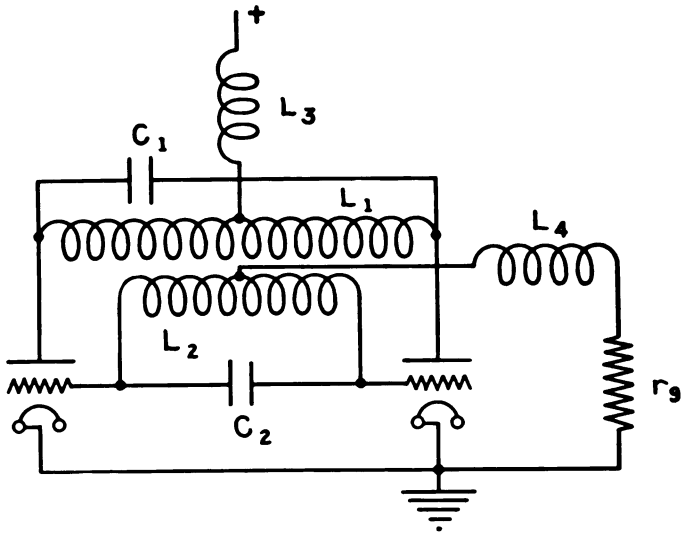

FIGURE 7

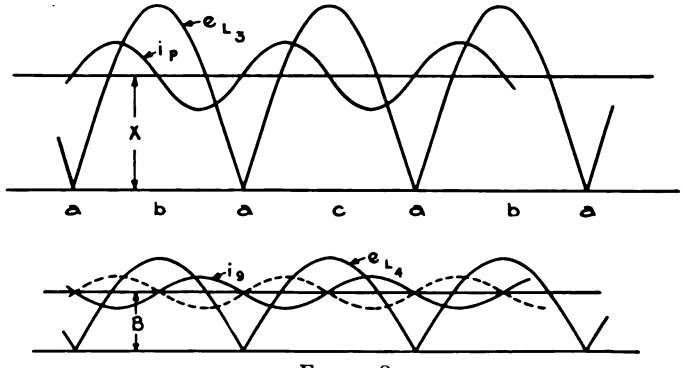

FIGURE 8

There is, therefore, nothing to prevent a rise in current except the inductance L_3. The current will therefore be rising at the points "a." To preserve the average, the counter emf. at times "b" and "c" is higher than the impressed potential X. The plate current i_p is, therefore, falling at "b" and "c."

By similar reasoning, e_{L_4} is the emf. at mid-point of L_2, due to the rectifier action of the grids. Since this potential is depended

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upon to make the grid current flow and since the potential is zero at points "a," the grid current is falling at these points and rising at points "b" and "c." The amplitude of the variations is dependent upon inductance L_r. The point of importance is that while i_p is rising, i_g is falling, and vice versa. In Figure 9 the tube losses for various values of grid voltage and plate current are shown. The corresponding plate voltage and grid current are found from the characteristic of Figure 6 from which Figure 9 was derived.

Suppose the required plate current for a given load to be 0.7 amperes, then the most economical grid voltage is 520, and the plate voltage is about 750, and the grid current about 0.225. The circuit

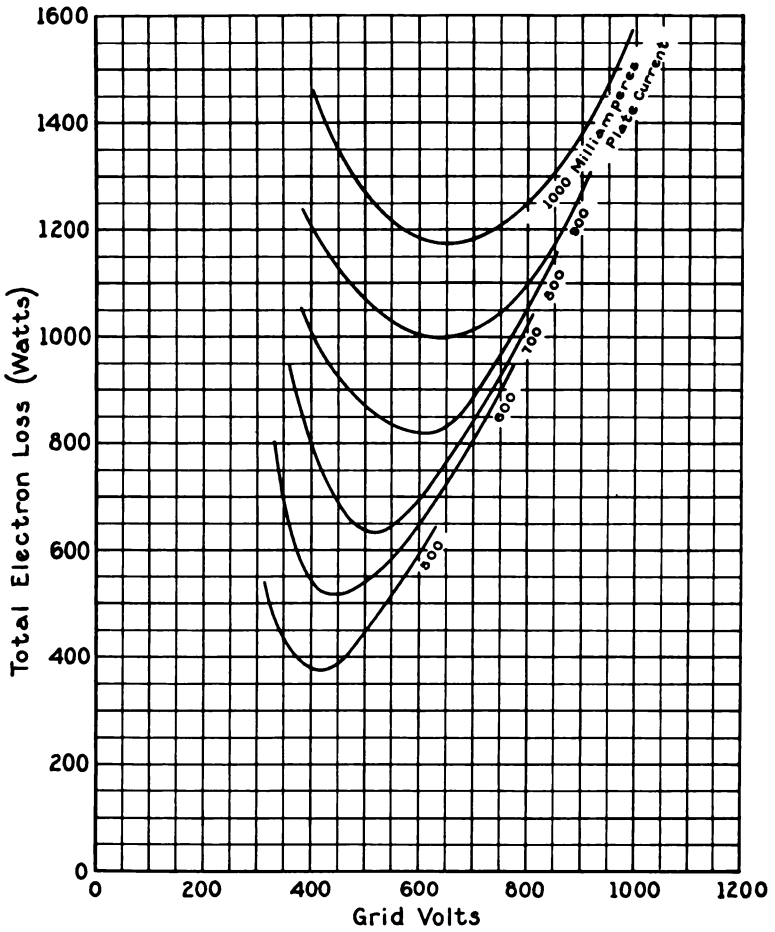

FIGURE 9—Space Charge Loss of the UV-206 Radiotron, the Characteristics of which are shown in Figure 6

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of Figure 7 would logically be designed so that these currents and voltages would be obtained. But if, for some reason, the currents vary, as shown in Figure 8, the plate current rising to, say 1 ampere, and the grid current falling to, say, 0.150, the plate potential must rise to nearly 1,200 volts, giving a loss of 1,275 watts as compared with 630 under the correct conditions.

On the corresponding down swing of plate current to 0.4 and up swing of grid current to 0.3, the loss falls to 270. The average loss is thus considerably higher if the values of L_3 and L_4 are such as to permit a material change in plate and grid currents during the cycle. It is therefore important with this circuit to keep variations in both plate and grid currents to a minimum.

With high frequency and high voltage, the design of inductances L_3 and L_4 becomes increasingly difficult. For instance, if the power factor of the circuit L_1 C_1 is 10 percent, the energy current is one-tenth of the circulating current. The direct current is roughly equal to this energy current, so that if the current ripple in the plate circuit is not to exceed 10 percent of the direct current, then the ohmic impedance of L_3 must be 100 times L_1. The lowest frequency impressed upon L_3 is twice the fundamental, so that L_3 must be 50 L_1 in henrys. At radio frequencies and with apparatus of reasonably small size, it is almost impossible to design the inductance L_3 so that its natural resonant frequency will be sufficiently above the frequency at which it is required to work. Such a circuit as that in Figure 7 is thus practically limited to a few hundreds of cycles.

Where conditions are such that the circuit of Figure 7 can be used, it is very desirable because it combines high efficiency with high output. The test of suitability is to find what inductance L_3 must have to keep the ripple within satisfactory limits and then see if it is possible to build such an inductance.

Rather than develop the design of the foregoing circuit in greater detail, a circuit suitable for a somewhat higher frequency will be described. The design of that circuit will include the work necessary to design such a circuit as that of Figure 7. Figure 10 is the circuit in question.

The arrangement in Figure 10 is precisely the same as that in Figure 7, except that L_3 and L_4 are magnetically coupled. By this means the voltages induced by the plate current ripple in L_3 may be utilized to cause a grid current similar to the plate current to flow. The grid current ripple i_g in Figure 8 is thus shifted 180° to the position shown dotted. In this way it is possible, even with a considerable ripple, to conform constantly to the conditions

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of minimum tube loss as obtained from Figures 6 and 9.

L_3 and L_4 are either actually connected thru a capacity as C_3, or the direct current power supply acts as such a condenser. If there should be no oscillations in C_1 L_1 and C_2 L_2, it is apparent that L_3 and L_4 may function as the plate and grid inductances of an ordinary Hartley circuit, the capacity coupling of which is furnished by the capacities between the coils representing L_1 and L_2 and such other parts of the main oscillating circuits as are exposed to one another. This circuit may support parasitic oscillations in which the two tubes operate in parallel and the frequency is the natural frequency of the circuits for this condition. The remedy for such parasitic oscillations is to have the fundamental circuit the lower decrement circuit of the two.

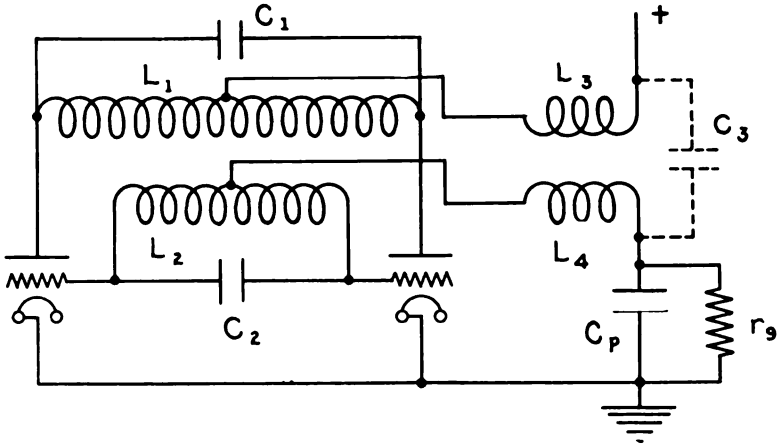

FIGURE 10

If this is not naturally the case, it may be made so by placing some resistance in series with C_p. The stored energy in the parasitic circuit with good design is small and its decrement is therefore fairly high. Another cause which tends to produce parasitic currents in any circuit is insufficient emission, and this must also be avoided.

The following sections of this paper are devoted to a discussion in detail of the design of the circuit Figure 10, with numerical example carried thru to illustrate how the various steps are taken.

RATING OF CIRCUIT

The rating of a circuit of the double-tube, square-wave type is very easy to predetermine. A given type of tube will be able

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safely to withstand the stresses caused by a certain limiting direct voltage and the temperatures due to the passage of a certain current thru it for half of each cycle. The product of this voltage and current is the input from the direct current source. The losses occurring in the tubes are readily determined, and those occurring in the coil systems may be estimated. Hence the output rating is quickly obtained.

Distribution of Energy Thruout Circuit

The ratio of the energy stored in an oscillating circuit to the energy dissipated per cycle is $\frac{1}{2\pi}$ times the ratio of reactive volt-amperes to watts. Hence, if it is desired to store twice as much energy in a circuit as is dissipated per cycle, the volt-amperes must be 4π times the watts. This has been found to be a very good criterion for the lower limit in the design of simple oscillating circuits, and it is doubtful if it is wise to make any reduction here; for altho energy is now introduced into the circuit in two pulses per cycle instead of one, it is more essential that the voltages thruout the circuit hold to a definite shape, and this is accomplished by utilizing the fairly pure sine wave obtained with moderate energy storage.

The volt-amperes given by the relation just discussed is the total for the circuit. Part of this is in the primary circuit, part is in the secondary circuit, and part is held by the two circuits in common. Another way of stating this is to say that the volt-amperes in the two condensers represent the total energy, and their sum is equal to the sum of the volt-amperes in the two coils, due to their self inductances plus a term involving both currents and the mutual inductance.

As the current required to commutate the grids is drawn from the condenser in the secondary circuit, it is obvious that this is one point in the circuit where it would be well to store an amount of energy quite large with respect to the local energy losses. Several circuits have been designed using 30-volt-amperes in the grid condenser for each watt lost at this point, and this seems to be a very good value. It certainly is generous, and at the same time does not ordinarily tie up too large a portion of the total energy.

Voltages in Plate and Grid Coils

The ratio of average to peak voltage for a double wave rectifier is $\frac{2}{\pi}$. The grid circuit is obviously an ordinary recti-

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fier, and the plate circuit is the same circuit reversed for the reception of energy at constant potential.

Due to the action of the grid choke, the voltage applied to the grid resistance is the theoretical average voltage minus the voltage drop in the grids themselves. Similarly, the plate choke takes the alternating voltage in the plate coil and presents it to the direct voltage source of energy as a steady counter electromotive force to be overcome. The input voltage must, of course, be equal to this counter emf., plus the plate drop.

Let E_p = plate drop under full load current.

E_g = grid drop under full load current.

E_1 = alternating voltage across primary coil (r. m. s. value).

E_2 = alternating current across secondary coil (r. m. s. value).

E_r = voltage applied to grid leak resistance (r_g).

E_c = voltage of constant potential source.

Then, remembering that only one-half the voltage of either coil is used at a time,

$$E_c = \tfrac{1}{2} \times \frac{2}{\pi} \times \sqrt{2}\, E_1 + E_p = \frac{\sqrt{2}}{\pi} E_1 + E_p \qquad (1)$$

and

$$E_r = \frac{\sqrt{2}}{\pi} E_2 - E_g \qquad (2)$$

It will be observed that these values are not entirely exact, as commutation of current from one tube to the other does not occur instantly at the time of zero coil voltage and other slight variations occur, but the values obtained will be found to be of quite sufficient accuracy for all practical needs.

The direct voltage of the supply source will, of course, be known, and this will determine E_1. The values of E_2 and E_r are not fixed until some assumption is made regarding the time required for the grid current to be transferred from one tube to the other. In general, this period will be short, but if it is made too short E_r will be found to be so high as to cause unwarranted losses in the grid resistance. The whole problem of commutation of the grid currents is quite elaborate and will be discussed later. For the present it will be sufficient to assume that commutation occupies a time interval represented by an angle θ to either side of the time when the voltage across the grid condenser is zero. See Figure 11. When commutation starts there is the drop E_g in one grid and zero drop across the other. The total voltage between grid terminals measured thru the grids is E_g.

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At the same time, this voltage as measured thru the grid condenser or coil is $\sqrt{2}\,E_2 \sin\theta$, and, as these two must be equal,

$$E_g = \sqrt{2}E_2 \sin\theta \tag{3}$$

As E_g will be known from the tube characteristics, the value of E_2 is determined when θ is assumed. As the volt-amperes in the primary and secondary condensers have already been determined, and the voltages are now known, the currents are easily obtained, thus giving the terminal conditions to be met by the plate and grid coils, which together form an air cored transformer.

Calculation of Coupling Between Plate and Grid Circuits

As the terminal conditions of the plate and grid coils are known, and if the coefficient of coupling be assumed, it will be found that all voltages and currents thruout this part of the circuit may be readily calculated. Since the maximum possible coefficient of coupling of such a transformer can be estimated from its volt-ampere and voltage ratings with a fair degree of accuracy for a given type of construction after a little experience, this is the logical factor to use as a basis of calculation. A high coefficient of coupling is desirable, as it makes for easy energy interchange between the circuits and prevents any appreciable phase differences. (This latter point is more fully covered in another section.) A very high coefficient of coupling, however, is obtained only at the expense of the insulation between coils. In general, it is usually found that a coefficient of coupling between 0.35 and 0.50 will give satisfactory results.

Proceeding to the method of calculation, let

$$\left.\begin{aligned}
P_1 = I_1^2\omega L_1 &\quad \text{be the wattless power in the primary, due to its own current and inductance.}\\
P_2 = I_2^2\omega L_2 &\quad \text{be the wattless power in the secondary, due to its own current and inductance.}\\
C_2 = I_1 I_2 \omega M_o &\quad \text{be the wattless power which is held between the two circuits.}\\
Q_1 = I_1^2/\omega C_1 &\quad \text{be the total primary volt-amperes.}\\
Q_2 = I_2^2/\omega C_2 &\quad \text{be the total secondary volt amperes.}
\end{aligned}\right\} \tag{4}$$

Since the only real power transferred thru the coupling is that utilized in the grids, and this is very small in comparison with the wattless volt-amperes involved, and since the coefficient of coupling is high, the primary and secondary currents may be as-

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sumed to be in phase in calculating their magnitudes; which will also make the primary and secondary voltages in phase.

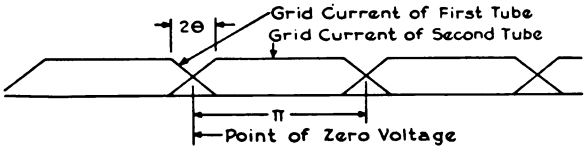

FIGURE 11

In order to fix the phase relationships definitely in mind, so as to avoid getting any term reversed in sign, let us refer to Figure 12. Current enters the two coils at a_1 and a_2 in the same

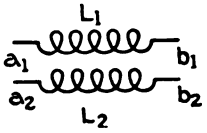

FIGURE 12

phase, and let us assume that at some instant the current in L_1 is developing its maximum effort to make a_1 positive with respect to b_1. At the same instant let the current in L_2 also be developing its maximum effort to the same end. Then both currents at the same instant are also exerting their peak effort to make a_2 positive with respect to b_2. Let these be the conditions assumed for positive values of P_1, P_2, C, and so on. Then

$$P_1+C=Q_1 \tag{5}$$

$$P_2+C=Q_2 \tag{6}$$

$$P_1P_3=\frac{C^2}{K^2} \tag{7}$$

where K is the coefficient of coupling. This last equation is obtained as follows:

$$M_o=K\sqrt{L_1 L_2} \qquad \text{(by definition)}$$

$$I_1 I_2 \omega M_o=K\sqrt{I_1^2 \omega L_1 \cdot I_2^2 \omega L_2}$$

$$C=K\sqrt{P_1 P_2}$$

$$P_1 P_3=\frac{C^2}{K^2}$$

Multiplying (5) by (6)

$$P_1 C+P_2 C+P_1 P_2+C^2=Q_1 Q_2 \tag{8}$$

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Substituting from (7)

$$P_1 C + P_2 C + \left(1 + \frac{1}{K^2}\right) C^2 = Q_1 Q_2 \tag{9}$$

Multiplying (5) by C

$$P_1 C = Q_1 C - C^2 \tag{10}$$

Multiplying (6) by C

$$P_2 C = Q_2 C - C^2 \tag{11}$$

Substituting (10) and (11) in (9)

$$(Q_1 + Q_2) C + \left(\frac{1}{K^2} - 1\right) C^2 = Q_1 Q_2$$

$$\left(\frac{1}{K^2} - 1\right) C^2 + (Q_1 + Q_2) C - Q_1 Q_2 = 0$$

$$C = \frac{-(Q_1 + Q_2) \pm \sqrt{(Q_1 + Q_2)^2 + 4 Q_1 Q_2 \left(\frac{1}{K^2} - 1\right)}}{2\left(\frac{1}{K^2} - 1\right)} \tag{12}$$

Inspection of the original equations will show that P_1 and P_2 may now be readily determined, thus giving the values of L_1, L_2, and M_o. As C_1 and C_2 (primary and secondary capacities) and r_g are easily derived from the results of the preceding paragraph, there remains only the load resistance, pass condenser for r_g, and inductance in the harmonic transformer to be determined.

If the circuit is loaded as shown in Figure 13, it is only necessary in calculating r_L to pick a resistance which will absorb the output power at the known current I_1.

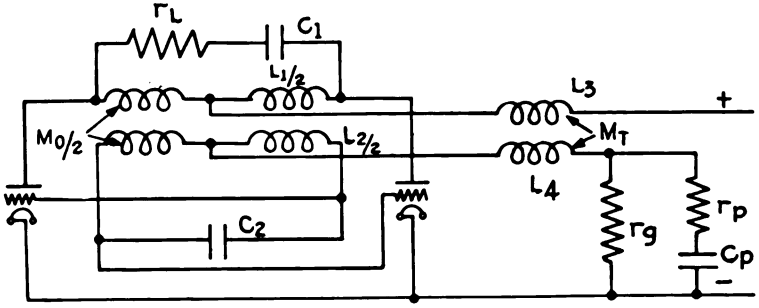

FIGURE 13—Diagram of Square Wave Circuit showing Nomenclature

The function of C_p is to pass the small amount of alternating current which gets thru the harmonic transformer so that it does not have to go thru r_g. The corresponding current which flows

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to the power supply thru L_3 will undoubtedly be short-circuited in this manner and it is desirable to have both paths of this same nature. The value of C_p is not critical; it must merely present a relatively low impedance to a double-frequency current of magnitude equal to a fraction of the grid current. The function of r_p in series with it is to discourage parasitic oscillations; its value is best found by trial.

The Harmonic Transformer

The function of the harmonic transformer is to pass the steady direct currents from the power source and grid leak resistance and to hold up the double frequency and its harmonic voltages which always appear wherever rectification takes place. Now it is impossible to do this without drawing some exciting current and it will be found that this exciting current must be quite large if bad internal tuning effects are to be avoided. If separate chokes are used in the two circuits the plate ripple will be increasing the plate current while the grid ripple will be decreasing the grid current. This is a condition giving bad additional losses in the tubes which could be reduced almost to zero if the two currents rose and fell together. This latter state may be obtained by coupling the two chokes together into a transformer and making the plate ripple current drive the grid current ripple in the right phase relation.

Before advancing any further, it is well to see what the magnitudes of the voltages are which it is proposed to hold back. The fundamental component of the ripple (that is the double output frequency component) for double wave rectifiers is $\frac{2}{3\pi}$ times the voltage across the outer terminals of the alternating current winding. Hence, neglecting all other components, and designating by E_3 and E_4 the voltages across L_3 and L_4 (see Figure 13),

$$E_3 = \frac{2}{3\pi} E_1 \quad \text{and} \quad E_4 = \frac{2}{3\pi} E_2 \tag{13}$$

As a further approximation assume that all the impedance presented to these voltages is that located in the harmonic transformer. The ripple currents will be taken as a certain percent of the direct currents. Their relative magnitude will depend on the particular circuit used. Some circuits due to relatively high direct currents and low voltages can be run with smaller ripples than others. The limit occurs due to internal resonance of the coils.

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We now have a transformer with given terminal conditions to be met, and the solution is the same as for the transformer in the oscillating circuit. Upon examination of the relative polarities of the voltages and currents, however, it is discovered if the previous nomenclature is held to, that the actual numerical value of the voltage applied to L_4 will be negative, whereas, before, everything had a positive numerical value. This has an important practical bearing on the problem as it puts a limit on the coefficient of coupling.

Let X_3 = reactance of coil L_3.
X_4 = reactance of coil L_4.
X_M = mutual reactance.
I_3 = alternating current thru L_3.
I_4 = alternating current thru L_4.

Then

$$I_3 X_3 + I_4 X_M = E_3 \tag{14}$$

$$I_4 X_4 + I_3 X_M = E_4 \tag{15}$$

$$X_3 X_4 = \frac{X_M^2}{K^2} \tag{16}$$

Multiplying (14) and (15) together and substituting from (16):

$$I_3^2 X_3 X_M + I_4^2 X_4 X_M + I_3 I_4 X_M^2 \left(1 + \frac{1}{K^2}\right) = E_3 E_4 \tag{17}$$

Multiplying (14) by $I_3 X_M$ and transposing

$$I_3^2 X_3 X_M = E_3 I_3 X_M - I_3 I_4 X_M^2 \tag{18}$$

Multiplying (15) by $I_4 X_M$ and transposing

$$I_4^2 X_4 X_M = E_4 I_4 X_M - I_3 I_4 X_M^2 \tag{19}$$

Substituting (18) and (19) in (17)

$$I_3 I_4 \left(\frac{1}{K^2} - 1\right) X_M^2 + (E_3 I_3 + E_4 I_4) X_M - E_3 E_4 = 0 \tag{20}$$

From which

$$X_M = \frac{-(E_3 I_3 + E_4 I_4) \pm \sqrt{(E_3 I_3 + E_4 I_4)^2 + 4 E_3 E_4 I_3 I_4 \left(\frac{1}{K^2} - 1\right)}}{2 I_3 I_4 \left(\frac{1}{K^2} - 1\right)} \tag{21}$$

In order that X_M may be a real number

$$(E_3 I_3 + E_4 I_4)^2 > -4 E_3 E_4 I_3 I_4 \left(\frac{1}{K^2} - 1\right)$$

$$E_3^2 I_3^2 + E_4^2 I_4^2 + 2 E_3 E_4 I_3 I_4 > -4 E_3 E_4 I_3 I_4 \left(\frac{1}{K^2} - 1\right)$$

$$E_3^2 I_3^2 + E_4^2 I_4^2 - 2 E_3 E_4 I_3 I_4 > -\frac{4 E_3 E_4 I_3 I_4}{K^2}$$

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Now the left-hand side is the square of a real number and hence positive. Also K^2 must be positive. Hence:

$$K^2 > \frac{-4 E_3 E_4 I_3 I_4}{E_3^2 I_3^2 + E_4^2 I_4^2 - 2 E_3 E_4 I_3 I_4} \qquad (22)$$

The limits on K are now known. If it is impossible or impracticable to build a set of coils having a coefficient of coupling within these values, it will be necessary to give new values to some of the assumed constants on which the circuit as been calculated. Changing the assumed ripples in the plate and grid currents will change only the actual values of X_3, X_4, and X_M, leaving their relative values unchanged. This is not apt to give any relief in the situation and earlier assumptions must be scrutinized, and of these the only ones which can be changed to advantage are those affecting the voltage in the grid circuit, for if this can be lowered, the requirements imposed on the harmonic transformer will be easier to meet as will be seen by examination of equation (22). As soon as a practical value of K is determined upon, the values of X_M, X_3, and X_4 follow from equations (21), (14), and (15), taking them in the order named.

Commutation of Grid Currents

Among the assumptions which were made earlier in this article was one to the effect that transfer of grid currents would be accomplished as tho this transfer had no effect on the voltage which would otherwise exist in the secondary circuit. It is now proposed to examine into the truth of this assumption. It is not practical to go very far into the investigation, however, so it will merely be assumed that during a short period of commutation, the current in the secondary coil remains constant, which will be very nearly the case, and the first assumption verified on this basis.

i_o = maximum of current flowing thru inductance L_2 and assume a constant current of this value to flow thru L_2 during commutation. (This is in addition to the direct components of the grid currents.)

i_1 = current in grid from which current is being transferred.

i_2 = current in grid to which current is being transferred.

i_g = steady value of grid current during the active part of the cycle. (This assumes complete smoothing by the harmonic transformer.)

r = resistance of two grids in series

Then it can be shown that

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$$i_2 = i_o\left(1-\varepsilon^{-\frac{t}{rC_2}}\right)+\frac{i_g}{2}\left(1-\varepsilon^{-\frac{t}{rC_2}}\right)=\left(i_o+\frac{i_g}{2}\right)\left(1-\varepsilon^{-\frac{t}{rC_2}}\right) \quad (23)$$

and

$$i_1 = -i_o\left(1-\varepsilon^{-\frac{t}{rC_2}}\right)+\frac{i_g}{2}\left(1+\varepsilon^{-\frac{t}{rC_2}}\right) \quad (24)$$

By setting i_2 equal to i_g it is possible to calculate the time at which transfer is completed, and in the usual case no appreciable difference is shown between this value and that given by the first assumption.

Losses in Tubes During Commutation

The loss in the plate circuit during commutation may be unduly increased by using too large a lap angle. This is a matter which probably cannot be calculated accurately, because of the lack of exact knowledge of many things which happen in an extremely short interval of time, but, if certain reasonable assumptions are made, at least the order of magnitude of the additional losses may be estimated. First let it be assumed that there is no phase displacement between the grid excitation and the plate voltage under which conditions commutation is effected by the grids. Then at the start and finish of the period of commutation, the tube which is carrying the load current should have only the voltage drop across it actually required to pass the current. The other tube will have the remainder of the voltage of the primary circuit impressed upon it. While it is extremely difficult to say exactly how the voltage will divide between the tubes during commutation the most likely division will be of the nature shown in Figure 14.

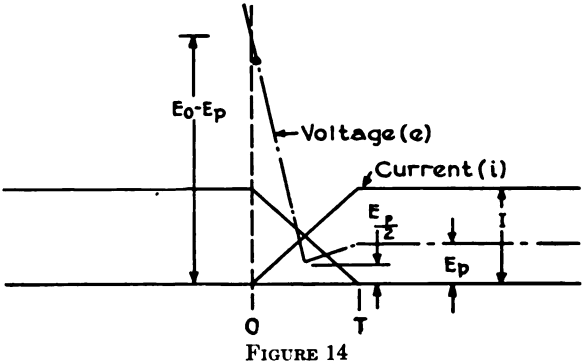

Figure 14

Let E_o be the voltage in the primary circuit at either end of the period of commutation.

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E_p be the steady plate drop while either tube is carrying all the load current.

e be the instantaneous voltage across the tube into which the current is being commutated.

i be the instantaneous current thru the same tube.

Then assume that

$$i = \frac{I}{T} t$$

$$e = (E_o - E_p) - \frac{2\left(E_o - \frac{3}{2} E_p\right)}{T} t \quad \text{for} \quad 0 < t < \frac{T}{2}$$

and

$$e = \frac{E_p}{T} t \quad \text{for} \quad \frac{T}{2} < t < T$$

Measuring t in degrees and calling the losses in both tubes W_1

$$W_1 = \frac{4}{360}\left\{\frac{(E_o - E_p)\, I}{T}\int_o^{\frac{T}{2}} t\, dt - \frac{2\,(E_o - \frac{3}{2} E_p)\, I}{T^2}\int_o^{\frac{T}{2}} t^2 dt + \frac{I\, E_p}{T^2}\int_{\frac{T}{2}}^{T} t^2\, dt\right\}$$

$$= \frac{I\, T}{2{,}160}\left\{E_o + 7\, E_p\right\}$$

Now let W_2 = watts lost in one tube during time $2\,T$ if it carried the steady current I. Then

$$W_2 = \frac{2\,T}{360} \times I \times E_p$$

When $W_1 = W_2$ there are no additional losses due to commutation

$$\frac{2\, E_p}{360} = \frac{E_o + 7\, E_p}{2{,}160}$$

or

$$E_o = 5\, E_p$$

and if $E_o < 5\, E_p$ it is unlikely that there will be any serious losses due to the commutation even tho our assumptions may not be obtained in practice.

Vector Diagram of Circuit

Up to this point the phase displacement between the grid excitation and plate voltage has been assumed to be zero. While it is usually very small it is often desirable to know its value. This can be determined by referring to Figure 15. The subcripts refer to the different parts of the circuit shown in Figure 13. The root-mean-square components of plate and grid currents are designated I_p and I_g.

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Starting with I_{L_2} we know that E_2 leads it by $90°+\theta_1$ in order that this coil may be a source of power for the secondary. I_{Cr} leads E_2 by 90° and I_g must have such a value that the vector sum of I_{L_2}, I_{C_2} and I_g is zero. E_2 consists of the voltage induced in L_2 by its own current ($I_{L_2}\ X_{L_2}$), the voltage induced in L_2 by the primary current ($I_{L_1}\ X_{M_o}$), and the resistance drop in L_2.

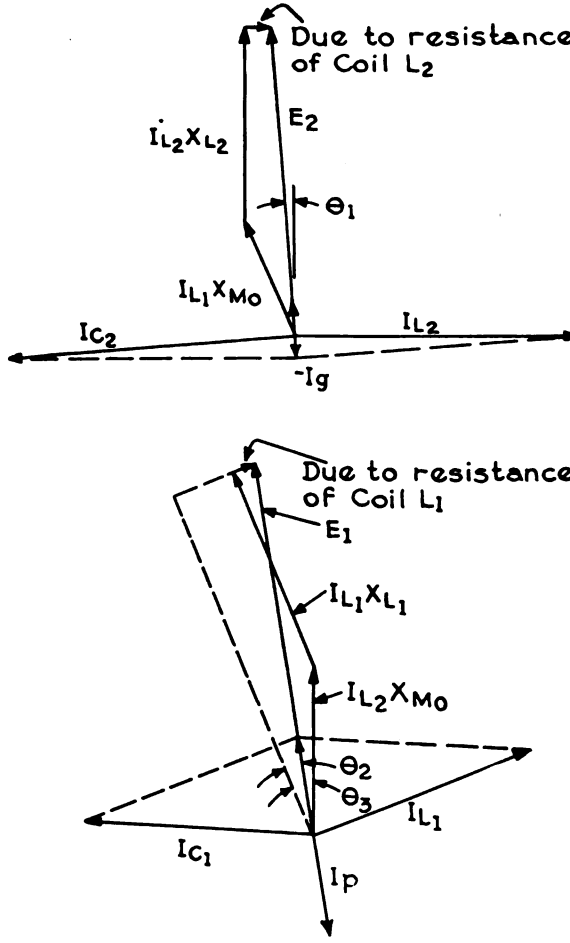

FIGURE 15—Vector Diagram of Circuit shown in Figure 13

In the primary circuit we have its own current I_{L_1} and the voltage induced in the primary coil by the secondary current ($I_{L_2}\ X_{M_o}$). Adding to the latter the voltage induced in the primary coil by its own current ($I_{L_1}\ X_{L_1}$) and the resistance drop in the same coil gives E_1, the primary voltage. I_p must then be

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180° out of phase with E_1 in order to represent input to the circuit and I_{C_1} must combine with I_{L_1} and I_p to give a vector sum of zero.

Let Q_1 = total primary reactive volt-amperes.

Q_2 = total secondary reactive volt-amperes.

C = reactive volt-amperes common to both circuits.

P_g = watts lost in grids.

Then if the resistance of the coils themselves be neglected

$$\left.\begin{aligned} \theta_1 &= tan^{-1}\frac{P_g}{Q_2} \\ \theta_2 &= tan^{-1}\frac{P_g}{Q_1} \\ \theta_3 &= tan^{-1}\frac{P_g}{C} \end{aligned}\right\} \qquad (25)$$

and the angle ϕ by which the grid excitation leads the plate voltage is given by

$$\phi = \theta_1 + \theta_2 - \theta_3 \qquad (26)$$

PRODUCTION OF LEAD ANGLE

In circuits using soft tubes, that is tubes which have control only to the extent that current flow may be held back if once stopped by external sources, it is necessary that the grid excitation shall lead the plate voltage in phase by such an amount that commutation is completed before the point of reversal of the commutating voltage is reached. This may be accomplished by placing a condenser in each plate lead bridged by a choke of high inductance to pass the direct component of plate current.

The displacement of voltages produced by such a device may be seen by referring to Figure 16. The tubes are here represented as a source of alternating current power feeding the oscillating circuit which, being tuned, may be represented as a pure resistance. The circuit is completed by the two phasing condensers. That part consisting of the oscillating circuit and phasing condensers is so drawn that the lengths and directions of the different impedances are representative of the voltages across the parts.

The grid voltage will be practically in phase with the voltage across the oscillating circuit; and any difference may be calculated by the method just shown. Hence the difference in phase between grid excitation and plate voltage is known as soon as the impedances are known; or, if the effective resistance of the oscillating circuit is known, the size of the condensers required to produce a given phase displacement may be calculated. Since

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the voltage across the oscillating circuit and the power output are known, its effective resistance is readily calculated.

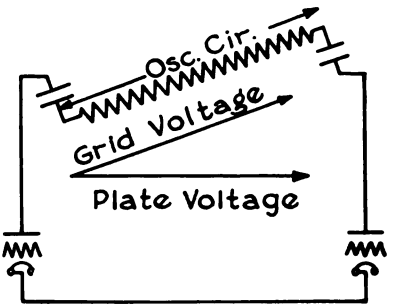

FIGURE 16

EXAMPLE ILLUSTRATING METHOD OF CALCULATION

The circuit to be calculated is to use two General Electric Company 50-watt tubes having "XL" filaments and is to draw 500 milliamperes from a 1,000 volt direct current source for input. The frequency is to be 12,000 cycles. Investigation of the tubes show that they will pass 500 milliamperes plate current with a plate drop of about 165 volts, a grid drop of about 125 volts, and a grid current of 85 millamperes. This gives:

Applied voltage (E_c)	$=1{,}000$ V.
Plate drop under full load current	$(E_p)=$ 165 V.
Grid drop under full load current	$(E_g)=$ 125 V.
Input current	$(I)=500$ ma.
Grid current	$(i_g)=$ 85 ma.
Input	$=$ 500 watts

Assuming 80 percent efficiency, the output will be 400 watts, and as the reactive power is to be 4π times the watts, there will be 5,000 volt-amperes in both circuits.

From equation (1) the voltage in the primary circuit (E_1) is

$$E_1=\frac{\pi}{\sqrt{2}}(E_c-E_p)$$

$$=\frac{\pi}{\sqrt{2}}(1{,}000-165)=1{,}850 \text{ } volts \text{ } (r.m.s.)$$

Let it be assumed that 10° will be sufficient time for commutation in the grid circuit. Then the r.m.s. voltage in the secondary circuit (E_2) is given by equation (3)

$$E_2=\frac{E_g}{\sqrt{2}\sin\theta}=\frac{125}{\sqrt{2}\times\sin 5^\circ}=1{,}015 \text{ } volts \text{ } (r.m.s.)$$

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and the voltage across the grid resistance (E_r) is given by equation (2).

$$E_r = \frac{\sqrt{2}}{\pi} E_2 - E_g$$

$$= \frac{\sqrt{2}}{\pi} \times 1{,}015 - 125 = 333 \text{ volts.}$$

At 85 ma., this represents 28 watts loss, and the impedance of the grid resistance is

$$r_g = \frac{333}{0.085} = 3{,}920 \text{ ohms.}$$

If it is assumed that 2,000 volt-amperes out of the total of 5,000 will be in the secondary circuit, the losses in the secondary circuit will be

Loss in grids $2 \times 125 \times 0.085$	= 21 watts
Loss in circuit due to resistance (at 1 percent power factor)	= 20 watts
Loss in r_g	= 28 watts
Total loss in secondary circuit	= 69 watts

This gives thirty volt-amperes in the secondary circuit for every watt lost, which appears to be a good value.

We can now show that commutation will occur in substantially the 10° assumed. This may be accomplished by the use of equation (23).

$$i_o = \sqrt{2} \times \frac{2{,}000}{E_2} = 2.79 \text{ amperes}$$

$$\frac{1}{2\pi f c_2} = \frac{E_2}{I_2} = \frac{1015}{0.707 \times 2.79}$$

from which $C_2 = 2.57 \times 10^{-8}$ microfarad

$$r = \frac{2E_g}{I_g} = \frac{2 \times 125}{0.085} = 2{,}940 \text{ ohms.}$$

Substituting in equation (23) for the finish of transfer.

$$i_2 = \left(i_o + \frac{i_g}{2}\right)\left(1 - \epsilon^{-\frac{t}{rC_2}}\right)$$

$$0.085 = (2.79 + 0.043)\left(1 - \epsilon^{-\frac{t}{rC_2}}\right)$$

from which

$$\frac{t}{rC_2} = 0.0304$$

$$t = 0.0304\, rC_2 = 0.0304 \times 2{,}940 \times 2.57 \times 10^{-8}$$

$$= 2.295 \times 10^{-6} \text{ second}$$

$$12{,}000 \text{ cycles} = 4.32 \times 10^{6} \text{ degrees/second.}$$

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Therefore, the period of transfer covers

$$2.295 \times 4.32 = 9.9°$$

It has already been decided to divide the reactive power in the circuit so that there are 3,000 volt-amperes in the primary (Q_1) and 2,000 volt-amperes in the secondary (Q_2). Hence if the coefficient of coupling between primary and secondary (K) be assumed, the value of the reactive power common to both circuits (C) may be obtained from equation (12).

$$C = \frac{-(Q_1+Q_2) \pm \sqrt{(Q_1+Q_2)^2 + 4Q_1Q_2\left(\frac{1}{K^2}-1\right)}}{2\left(\frac{1}{K^2}-1\right)}$$

and if $K = 0.50$, then $\left(\frac{1}{K^2}-1\right) = 3$, and

$$C = \frac{-5{,}000 \pm \sqrt{5{,}000^2 + 4 \times 3{,}000 \times 2{,}000 \times 3}}{2 \times 3}$$

$= +808$ or $-2{,}475$; and the positive value is the one desired in this case.

Since P_g the loss due to the grids is 69 watts, the phase displacement between grid and plate voltages may be calculated by equations (25) and (26).

$$\theta_1 = tan^{-1}\frac{P_g}{Q_2} = tan^{-1}\frac{69}{2{,}000} = 2°$$

$$\theta_2 = tan^{-1}\frac{P_g}{Q_1} = tan^{-1}\frac{69}{3{,}000} = 1°20'$$

$$\theta_3 = tan^{-1}\frac{P_g}{C} = tan^{-1}\frac{69}{808} = 4°55'$$

$$\phi = \theta_1 + \theta_2 - \theta_3 = -(1°\,35')$$

Calculation of Constants of Main Oscillating Circuit

E_1, the primary voltage, has been shown to be 1,850 volts r.m.s.

E_2, the secondary voltage, has been shown to be 1,015 volts r.m.s. Since there are 3,000 volt-amperes in the primary, the current is

$$I_1 = \frac{3{,}000}{E_1} = \frac{3{,}000}{1{,}850} = 1{,}622 \text{ ampere r.m.s.},$$

and since there are 2,000 volt-amperes in the secondary, the current here is

$$I_2 = \frac{2{,}000}{E_2} = \frac{2{,}000}{1{,}015} = 1.97 \text{ ampere r.m.s.}$$

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There are 808 volt-amperes common to both circuits oi

$$\omega M_o I_1 I_2 = 808$$

$\omega = 75{,}400$ for 12,000 cycles, hence

$$\omega M_o = \frac{808}{1.622 \times 1.97} = 253 \text{ ohms}$$

$$M_o = \frac{253}{75{,}400} = 3.36 \text{ milli-henrys.}$$

The volt-amperes in the primary coil due to its self inductance are

$$\omega L_1 I_1^2 = 3{,}000 - 808 = 2{,}192$$

$$\omega L_1 = \frac{2{,}192}{(1.622)^2} = 833 \text{ ohms.}$$

$$L_1 = \frac{833}{75{,}400} = 11.04 \text{ milli-henrys.}$$

The volt-amperes in the secondary coil due to its self-inductance are

$$\omega L_2 I_2^2 = 2{,}000 - 808 = 1{,}192$$

$$\omega L_2 = \frac{1{,}192}{(1.97)^2} = 307 \text{ ohms}$$

$$L_2 = \frac{307}{75{,}400} = 4.08 \text{ milli-henrys.}$$

The volt-amperes in the primary capacity are

$$\frac{I_1^2}{\omega C_1} = 3{,}000$$

$$\frac{1}{\omega C_1} = \frac{3{,}000}{(1.622)^2} = 1{,}140 \text{ ohms.}$$

$$C_1 = 0.01163 \text{ microfarad.}$$

The volt-amperes in the secondary capacity are

$$\frac{I_2^2}{\omega C_2} = 2{,}000$$

$$\frac{1}{\omega C_2} = \frac{2{,}000}{(1.97)^2} = 516 \text{ ohms.}$$

$$C_2 = 0.0257 \text{ microfarard.}$$

The voltages across the harmonic transformer are given by equations (13).

$$E_3 = \frac{2}{3\pi} E_1 = \frac{2}{3\pi} \times 1{,}850 = 394 \text{ volts r.m.s}$$

$$= 557 \text{ volts (peak)}$$

$$E_4 = \frac{2}{3\pi} E_2 = \frac{2}{3\pi} \times 1{,}015 = 215 \text{ volts r.m.s.}$$

$$= 306 \text{ volts (peak)}$$

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and this latter will be negative in sign under the assumed system of nomenclature.

Let it be assumed that these voltages cause ripple currents to flow (I_3 and I_4) which have peaks equal to 20 percent of the direct currents. Hence

$I_3 = 0.2 \times I = 0.2 \times 0.5 = 0.1$ *ampere (peak)* $= 0.0707$ *ampere r.m.s.*

$I_4 = 0.2 \times i_g = 0.2 \times 0.085 = 0.017$ *ampere (peak)* $= 0.012$ *ampere r.m.s.*

Using equation (22) to find the minimum possible coupling gives

$$K^2 > \frac{-4}{\frac{E_3 I_3}{E_4 I_4} + \frac{E_4 I_4}{E_3 I_3} - 2}$$

$$K^2 > \frac{-4}{-\frac{394 \times .0707}{215 \times .012} - \frac{215 \times .012}{394 \times .0707} - 2}$$

$$K^2 > 0.313.$$

This gives a value of 0.56 as the minimum for K, and substituting this value in equation (21) gives

$$X_M = \frac{-(E_3 I_3 + E_4 I_4) \pm \sqrt{(E_3 I_3 + E_4 I_4)^2 + 4 E_3 E_4 I_3 I_4 \left(\frac{1}{K^2} - 1\right)}}{2 I_3 I_4 \left(\frac{1}{K^2} - 1\right)}$$

$$= \frac{-(394 \times 0.0707 - 215 \times 0.012) \pm 0}{2 \times 0.0707 \times 0.012 \times (3.19 - 1)} = -6{,}780 \text{ ohms.}$$

Knowing X_M, the values of X_3 and X_4, the reactances of coils L_3 and L_4, are obtained from equations (14) and (15)

$$X_3 = \frac{E_3 - I_4 X_M}{I_3}$$

$$= \frac{394 + 0.012 \times 6{,}780}{0.0707} = 6{,}720 \text{ ohms.}$$

$$X_4 = \frac{E_4 - I_3 X_M}{I_4}$$

$$= \frac{-215 + 0.0707 \times 6{,}780}{0.012} = 21{,}900 \text{ ohms.}$$

At 24,000 cycles (150,800 radians/seconds) these impedances give

$M_T = -45$ *milli-henrys*

$L_3 = 44.6$ *milli-henrys*

$L_4 = 145.3$ *milli-henrys*

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Experimental Performance of Small Circuit

In order to show the probable variations between actual performance and performance as calculated, the circuit which has just been analyzed was set up in the laboratory. The tubes had been selected and tested before the calculations were started, and the actual and calculated values of the other parts of the circuit are given in Table I.

TABLE I

12,000 ~ Square Wave Circuit Using 2-50 Watt "XL" Filament Tubes

Part	Calculated Value	Actual Value	Resistance
C_1	0.01163 mfd.	0.012 mfd.	.....
C_2	0.0257 mfd.	0.026 mfd.	.....
C_p	.....	2.0 mfd.	.....
L_1	11.04 milli-henrys	11.2 mh.	6 ohms
L_2	4.08 milli-henrys	4.0 mh.	3.33 ohms
L_3	44.6 milli-henrys	44.67 mh.	36.5 ohms
L_4	145.3 milli-henrys	146.6 mh.	119 ohms
M_o	+3.36 milli-henrys	+3.43 mh.	.....
M_T	−45 milli-henrys	−44.9 mh.	.....
r_L	.....	.....	.....
r_g	3,920 ohms	4,100 ohms	.....
r_p	.....	500 ohms	.....

The harmonic transformer tuned with its internal capacity plus the capacity between the primary and secondary coils of the oscillating circuit at about 78,000 cycles.

The performance of the circuit is shown in Figure 17. The output as measured along the axis of abscissas is the actual useful output in the resistance r_L. Now the calculations were based on 500 watts input, and from the experimental curves this corresponds to 329 watts output. Drawing a vertical line thru this value, and inserting on it the calculated values for the different currents and the tube losses shows that the performance is practically as calculated.

Research Laboratory,
General Electric Company,
Schenectady, New York.

August 21, 1923.

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and this latter will be negative in sign under the assumed of nomenclature.

Let it be assumed that these voltages cause ripple cu flow (I_3 and I_4) which have peaks equal to 20 percent of t currents. Hence

$I_3 = 0.2 \times I = 0.2 \times 0.5 = 0.1$ *ampere (peak)* $= 0.0707$ *am*

$I_4 = 0.2 \times i_g = 0.2 \times 0.085 = 0.017$ *ampere (peak)* $= 0.0$
r.m.s.

Using equation (22) to find the minimum possible co

$$K^2 > \frac{-4}{\frac{E_3 I_3}{E_4 I_4} + \frac{E_4 I_4}{E_3 I_3} - 2}$$

$$K^2 > \frac{-4}{-\frac{394 \times .0707}{215 \times .012} - \frac{215 \times .012}{394 \times .0707} - 2}$$

$$K^2 > 0.313.$$

This gives a value of 0.56 as the minimum f tuting this value in equation (21) gives

$$X_M = \frac{-(E_3 I_3 + E_4 I_4) \pm \sqrt{(E_3 I_3 + E_4 I_4)^2 + 4 I}}{2 I_3 I_4 \left(\frac{1}{K^2} - 1\right)}$$

$$= \frac{-(394 \times 0.0707 - 215 \times 0.012) \pm 0}{2 \times 0.0707 \times 0.012 \times (3.19 - 1)} =$$

Knowing X_M, the values of X_3 and X_4, t L_3 and L_4, are obtained from equations (14

$$X_3 = \frac{E_3 - I_4 X_M}{I_3}$$

$$= \frac{394 + 0.012 \times 6{,}780}{0.0707} = 6.7$$

$$X_4 = \frac{E_4 - I_3 X_M}{I_4}$$

$$= \frac{-215 + 0.0707 \times 6{,}780}{}$$

At 24,000 cycles give

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Total Range and Accu-
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includes the antenna capac-
ill be considered first.
circuit tuned with a variable
nd minimum capacity of C_{max}

is designated C_a, and the maxi-
hs by λ_{max} and λ_{min}.

$$= \frac{C_{max}}{C_{min}}$$

$$S = \frac{C_{max}}{C_a}$$

$$T = \frac{\lambda_{max}}{\lambda_{min}}$$

e generally known at the start, except
es the receiver is supposed to work with
which case the writer generally uses the
e practical limit.
of taps on the inductance necessary can
m the formula:

$$N = \frac{log\ T}{log\ 0.90\sqrt{\frac{R+S}{1+S}}}$$

Editor, August 23, 1924.

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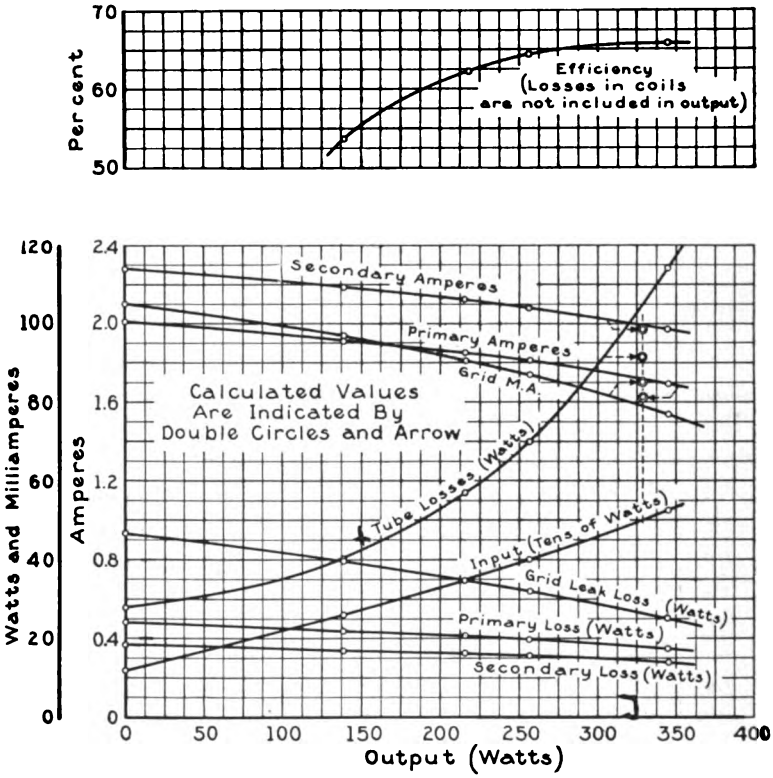

FIGURE 17—Performance of a Square-Wave Circuit with Constants given in Table I at 1,000 volts direct potential

SUMMARY: A modified push-pull circuit is described in which approximately square waves of current are passed thru the two tubes alternately. High efficiency is obtained by superimposing upon the grid circuit a voltage proportional to current variations in the plate circuit. The circuit gives high output combined with high efficiency without shortening the life of the tubes.

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Discussion* On "On the Calculation of the Inductances and Capacities for a Multi-Range . . . Series of Tuned Circuits . . . "—By J. Erskine Murray

By

Ralph R. Batcher

Referring to the paper from J. Erskine-Murray in the August, 1924, issue, "On the Calculation of the Inductances and Capacities for a Multi-Range or Other Consecutive Series of Tuned Transmitting or Receiving Circuits, the Total Range and Accuracy Required Being Given," the writer has devised another method which has proven of practical value for several years, and which seems considerably simpler than the method outlined.

Since many of the present-day circuits require the design of a multi-range oscillating circuit which includes the antenna capacity and inductance, this problem will be considered first.

The first case will be that of a circuit tuned with a variable condenser having a maximum and minimum capacity of C_{max} and C_{min}, respectively.

The capacity of the antenna is designated C_a, and the maximum and minimum wave lengths by λ_{max} and λ_{min}.

$$R = \frac{C_{max}}{C_{min}}$$

$$S = \frac{C_{max}}{C_a}$$

$$T = \frac{\lambda_{max}}{\lambda_{min}}$$

All of these values are generally known at the start, except possibly S. In many cases the receiver is supposed to work with a number of aerials, in which case the writer generally uses the value $S = 4$, which is the practical limit.

The total number of taps on the inductance necessary can then be computed from the formula:

$$N = \frac{\log T}{\log 0.90\sqrt{\frac{R+S}{1+S}}}$$

*Received by the Editor, August 23, 1924.

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where the factor 0.90 insures a 10 percent overlap. The derivation is not difficult, but will not be given here.

Having determined the number of taps (N) required, the location of each is given from the formula:

The last tap $= \text{Total inductance} = L$ (max.)
Next to the last tap $= L_{max} \div x$
Second from last $= L_{max} \div x^2$
Third from last $= L_{max} \div x^3$
nth from last $= L_{max} \div x^n$.

The value for L_{max} can be determined from

$$L_{max} = \lambda^2 \frac{1+S}{3{,}552\, C_{max}}$$

In some cases in computing the two lowest taps, it may be desirable to consider the inductance of the aerial. In most cases sufficient accuracy will be obtained if the equivalent concentrated inductance of the aerial is subtracted from the values of inductance obtained by the above formulas. A number of cases have been solved using a more exact formula, but the additional accuracy obtained rarely warranted the extra amount of work. The simpler case, of where the antenna is not included in the oscillating circuit, corresponding to the problem considered by Mr. Erskine-Murray, is solved in a similar manner.

Such a circuit might, for example, be a secondary circuit of which the above coil is the primary or a wavemeter. Naturally the same wave length bands must be covered.

$$\text{Then} \quad \frac{L_{sec}}{L_{pri}} = \frac{M}{S+1} \quad \text{where}$$

M is the ratio of maximum condenser capacities of each circuit, that is,

$$M = \frac{C_{pri}}{C_{sec}}$$

The number of taps is given by the formula

$$N_s = \frac{log\ T}{log\ 0.90 \sqrt{\frac{R\,D+1}{D+1}}}$$

and the locations of the taps are determined in the same way described except that now

$$x_s = 0.90 \sqrt{\frac{R\,D+1}{D+1}}$$

where R is the condenser capacity ratio as before, and D is the

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ratio of the minimum capacity of the condenser to the distributed capacity of the attached circuit, such as that due to leads, coils, vacuum tubes, and so on.

Using this method for locating the taps, the wave lengths will only be found in one location (except at the ends) and the minimum number of taps possible is used, so that tuning is much easier. In the broadcast receiver field, it may be mentioned, the use of the "straight line frequency" condenser simplifies tuning to a remarkable degree since all frequency bands are separated by an equal amount of dial displacement. The accuracy of tuning, outlined by Mr. Erskine-Murray, is constant over the whole range.

August 8, 1924.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

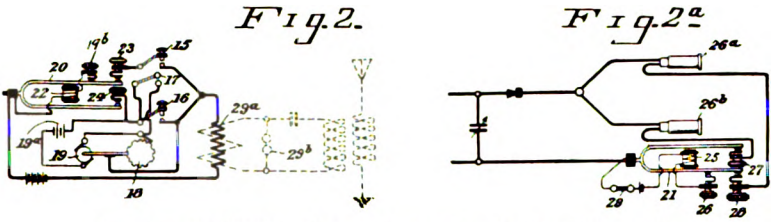
ISSUED JUNE 24, 1924—AUGUST 19, 1924

BY

JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, D. C.)

1,505,055—A. R. Molins, filed May 19, 1922, issued August 12, 1924.



NUMBER 1,505,055—Secret System for Radio Telegraphy

SECRET SYSTEM FOR RADIO TELEGRAPHY, wherein a tuning fork is provided at both the transmitting and receiving stations and arranged to vibrate in synchronism to close sets of contacts connected in different circuits whereby one series of signals may be radiated between the spaces of another series of signals. The messages are therefore transmitted in mixed relation and separated by a synchronized device at the distant receiving station.

The following patent of interest issued in 1922 has been added to this digest:

1,438,828—H. W. Houck, filed March 29, 1920, issued December 12, 1922.

METHOD AND APPARATUS FOR SELECTIVELY TRANSFERRING ELECTRICAL OSCILLATORY ENERGY, of any frequency or frequencies lying within a continuous band of frequencies from one electrical system to another. This patent shows an electron tube

*Received by the Editor, September 12, 1924. Readers will note that the first patent, 1,505,055, was issued August 12, 1924, and inadvertently omitted from an earlier digest.

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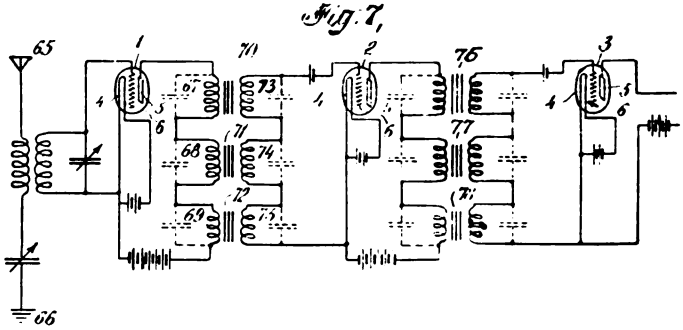
amplifier in which the input and output circuits of the several tubes are coupled by means of a series of oscillatory circuits tuned to different frequencies. Each circuit is resonant to a different frequency so that the circuit has a highly efficient collective range of resonance which includes the band of frequencies.



NUMBER 1,438,828—Method and Apparatus for Selectively Transferring Electrical Oscillatory Energy

1,498,536—S. N. Baruch, filed August 2, 1920, issued June 24, 1924.

APPARATUS FOR PRODUCING ELECTRICAL OSCILLATIONS comprising an arc generator where the electrodes are disposed on a horizontal plane between a vertically disposed magnetic field arranged to produce a field transversely of the arc. The lower magnetic pole is provided with a cup-shaped extension which carries a quantity of mercury for the production of a mercury vapor in the arc chamber.

1,498,898—A. L. Young, filed July 25, 1922, issued June 24, 1924.

SPREADER FOR ELECTRICAL WIRES formed in a cage type of antenna. A circular spreader is provided for spacing the antenna wires of a cage antenna in a cylindrical frame-like body. A detachable binding wire is passed around the circular spreader and is drawn taut to secure the wires with reference to the spreader.

1,499,310—J. Bethenod, filed August 2, 1921, issued June 24, 1924.

CALLING SYSTEM FOR RADIO SIGNALING, where a thermo-electric element is connected to the receiving circuit and heated by the incoming signaling energy to generate a direct current for operating a ballistic galvanometer. The galvanometer has con-

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siderable inertia and is unresponsive to short impulses. In order to operate the calling device, an extended signal is transmitted over such a period of time that the thermo-electric element becomes sufficiently heated to produce the necessary continuous energy for operating the galvanometer, which in turn closes the local calling circuit.

1,499,331—M. C. Batsel, filed December 11, 1922, issued July 1, 1924. Assigned to Westinghouse Electric and Manufacturing Company of Pittsburgh, Pennsylvania.

RADIO RECEIVING CIRCUIT employing the feed-back principle where the input and output circuits of an electron tube each are provided with variable inductors for providing the feed-back coupling there between. Additional inductance is provided in each of the circuits for establishing such additional feed-back coupling that the degree of regeneration is substantially independent of all adjustments of the inductor in the input circuit of the tube. The circuit arrangement provides a compact radio receiving set.

1,499,403—W. H. Priess, filed December 5, 1924, issued July 1, 1924. Assigned, Wireless Specialty Apparatus Company, Boston, Massachusetts.

ELECTRICAL CONDENSER for high power operation having a plurality of stacks of sheets carried within a metallic casing by which the stacks are connected in series. The opposite terminals for the condenser extend thru the metallic casing and are well insulated therefrom.

1,499,404—William H. Priess, filed December 5, 1921, issued July 1, 1924. Assigned, Wireless Specialty Apparatus Company, Boston, Massachusetts.

ELECTRICAL CONDENSER for use in radio transmission wherein maximum heat radiation from a plurality of stacks may be secured so that the mica dielectric may be operated at a desirable point on its phase angle temperature curve. To allow for rapid heat conduction and radiation the condenser is arranged in a plurality of relatively short stacks and maintained under pressure between heat radiating plates.

1,499,865—H. E. Hallborg, filed September 1, 1923, issued July 1, 1924.

HIGH SPEED SIGNALING SYSTEM, wherein a plurality of sets

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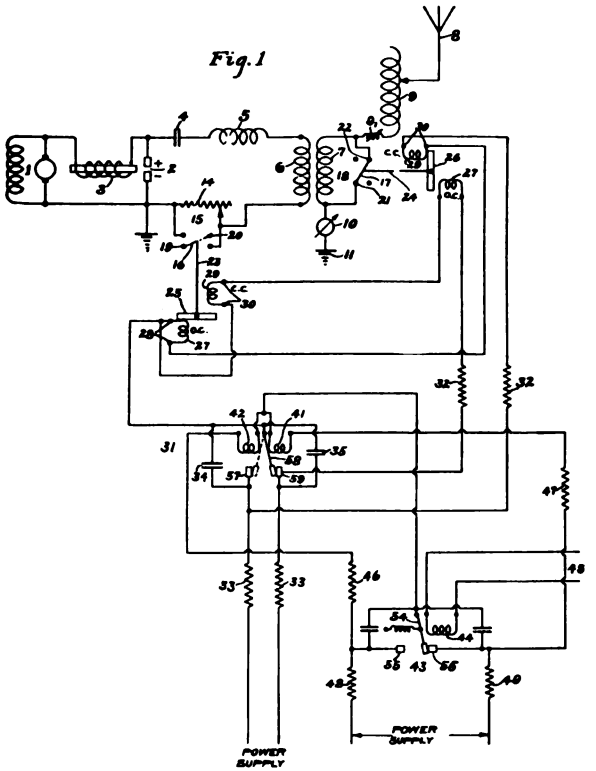
of keys are employed with an armature for actuating each of the keys with negligible electrical time lag. A plurality of opening coils and a plurality of closing coils are arranged to actuate each of the armatures. The electrical inertia of the control circuit which connects the opening and closing coils is extremely small, permitting high speed operation of the key.



NUMBER 1,499,865—High Speed Signaling System

1,500,476—F. K. Vreeland, filed July 28, 1920, issued July 8, 1924.

INTERFERENCE PREVENTION IN RADIO RECEPTION, having a pair of energy-collecting systems electrically coupled one with the other. The system is tuned to the desired signaling frequency and then another current of interfering frequency whose effect on the receiving system is opposite to any interfering effect which may be present when the signal is produced. These opposite effects are balanced in intensity and phase so that the resultant effect on the receiver is nil. This balancing of the interfering

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signals is accomplished without perceptible reduction in the signal strength of the energy desired to be received.

1,500,528—F. F. Rathbun, filed July 7, 1922, issued July 8, 1924.

VARIABLE PLATE ELECTRIC CONDENSER having cast-in stationary and movable plates for facilitating the protection of the instrument. The stationary plates are supported at three points about the periphery thereof by cast-metal poured into slotted tubular members. The movable plates are similarly supported by a slotted tubular member in which molten metal is poured over the plates.

1,501,104—L. Epenschied, filed September 26, 1919, issued July 15, 1924. Assigned to American Telephone and Telegraph Company, of New York.

MULTIPLEX RADIO SYSTEM, having a plurality of transmitting channels and a corresponding number of receiving channels. Different frequencies are superposed on the different channels, and the effect of these frequencies one upon the other is substantially eliminated by circuits provided in the different channels.

1,501,132—P. B. Murphy, filed May 4, 1921, issued July 15, 1924. Assigned to Western Electric Company, of New York.

SIGNALING SYSTEM employing a carrier wave having an alternate frequency source for modulating the radio frequency wave. According to this invention, at the transmitting station two oscillation generators each designed to supply a current of pre-determined frequency are alternately paralyzed at a lower frequency by a control oscillator, and at the receiving station the signal translating device is supplied with impulses alternating at double the frequency of the control oscillator. The several frequencies are employed at the receiver for controlling the receiving device.

1,501,240—K. Rottgardt, filed August 26, 1921, issued July 15, 1924.

MANUFACTURE AND OPERATION OF VACUUM TUBES, where the electrodes are freed of occluded gases by heating the elements to incandescence and introducing a hydrocarbon vapor into the tube and then evacuating the tube. In this manner the highest possible vacuum is secured and the last remnants of gas are removed from the metal.

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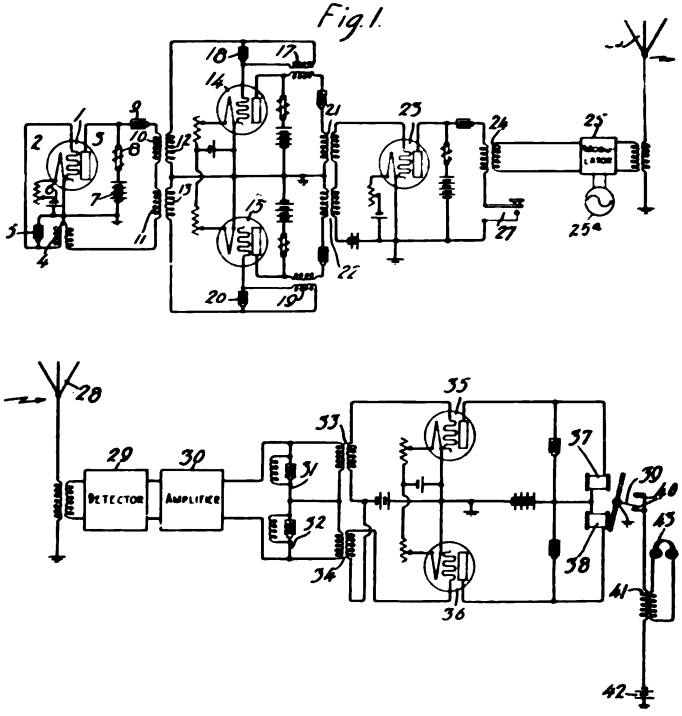

NUMBER 1,501,132—Signaling System

1,501,402—G. Jacobs, filed September 22, 1922, issued July 15, 1924.

VARIABLE CONDENSER, having a broad adjustment member and a fine adjustment member where the rotary plates are set in a tubular insulator having circumferential slots and grooves therein. The stationary plates project into the grooves and the rotary plates may be inter-leaved between the stationary plates.

1,501,543—L. A. Hammarlund, filed November 28, 1922, issued July 15, 1924. Assigned to Hammarlund Manufacturing Company, of New York.

VARIABLE CONDENSER, having broad adjustment and fine adjustment means consisting of a rock shaft on which the movable plates are mounted and a swinging arm operated by a cam for frictionally engaging the rock shaft and moving the same through small angles. The rotary plates are inter-leaved between the stationary plates and the relative position changed both by means of the rock shaft and the secondary swinging arm.

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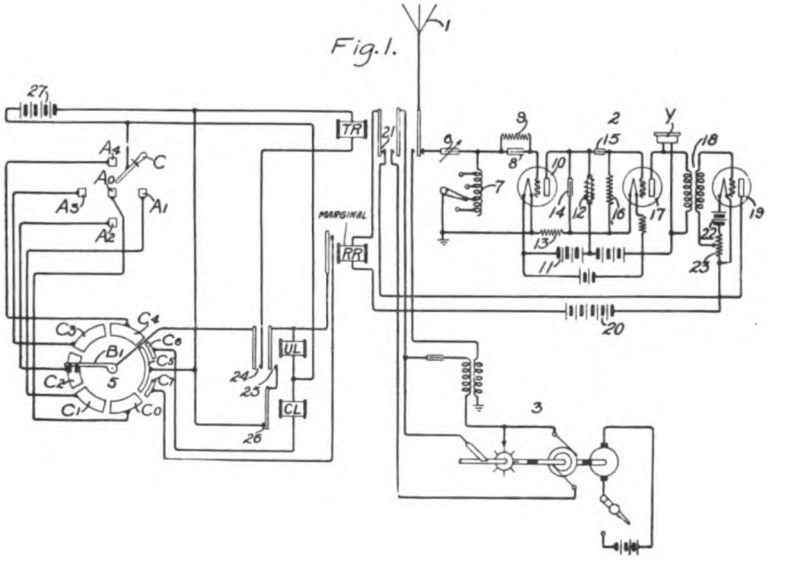
1,501,664—A. Hund, filed September 18, 1924, issued July 15, 1924. Assigned to General Electric Company, of New York.

RADIO SIGNALING SYSTEM for telephone transmission where modulation is secured by means of an electron tube circuit connected across points in an inductance series connected in the antenna. The electron tube circuit is controlled by a telephone transmitter for varying the amplitude of the transmitted signals.

1,501,683—A. A. Oswald, filed December 6, 1920, issued July 15, 1924. Assigned to Western Electric Company, of New York.

REMOTE CONTROL SYSTEM, having a tramsmitting device arranged to transmit impulses at timed intervals and a receiving device arranged to respond only to impulses received at definite instants and of a duration between two fixed limits. The system is substantially secret whereby a particular control may be locked at a receiver upon the transmission of a definite signal.

1,501,684—A. A. Oswald, filed December 7, 1920, issued July 15, 1924. Assigned to Western Electric Company, of New York.



NUMBER 1,501,684—Remote Control System

REMOTE CONTROL SYSTEM, wherein a rotary distributor at both a radio transmitter and receiver are maintained in synchronism by synchronizing impulses sent out from the transmitting station. A circuit arrangement is provided which pre-

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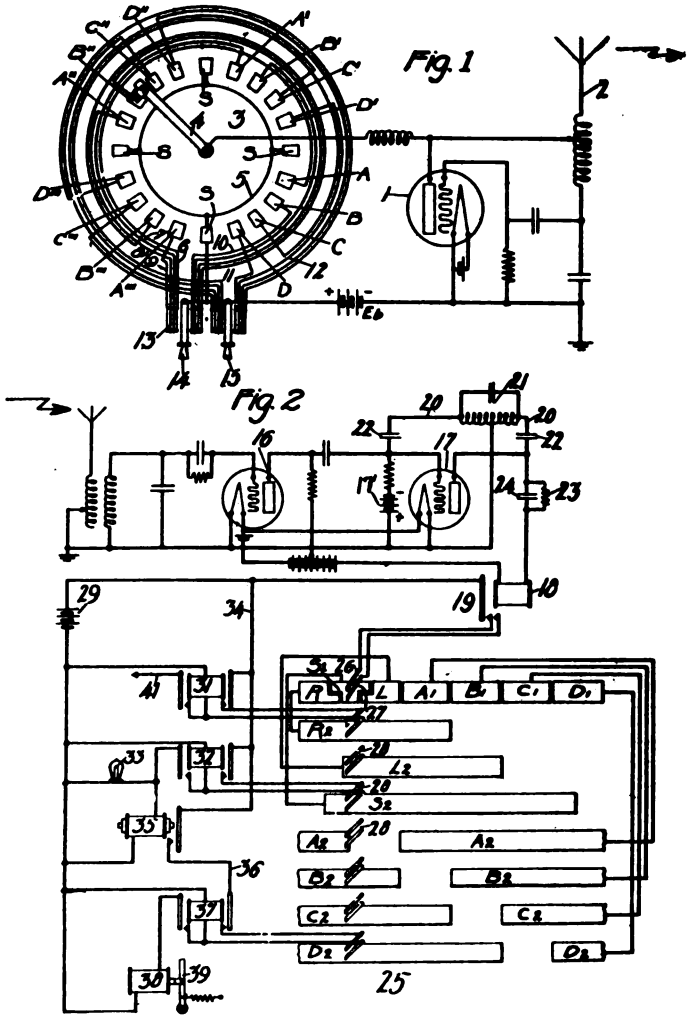
vents operation of the controlled devices except when the moving mechanism is in synchronism.

1,501,711—R. A. Heising, filed September 29, 1919, issued July 15, 1924. Assigned to Western Electric Company.



NUMBER 1,501,711—Distant Control System

DISTANT CONTROL SYSTEM, in which a group of oscillations is transmitted at a selected period within a timing cycle to determine the particular one of a plurality of distant controls to be

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actuated. At the receiving station the apparatus is maintained in proper phase relation to the transmitter and selected controls actuated.

1,501,729—E. O. Scriven, filed July 14, 1917, issued July 15, 1924. Assigned to Western Electric Company, of New York.

OSCILLATING GENERATOR having a circuit connected therewith for suppressing the undesired oscillations from the source. A resistance is connected between the source and the filter circuit which suppresses the undesired oscillations to prevent reaction between the filter and the source.

1,501,735—W. C. White, filed June 5, 1915, issued July 15, 1924. Assigned to General Electric Company, of New York.

RADIO SIGNALING SYSTEM, wherein the transmitter radiates a large amount of energy with a large variation in amplitude of the transmitted energy. An election tube circuit is provided having its cathode and anode connected across a high potential point in the antenna and the ground. Continuous wave energy is supplied to the antenna and the signals are produced by variation in the grid circuit.

1,501,830—E. F. W. Alexanderson, filed April 30, 1914, issued July 15, 1924. Assigned to General Electric Company.

RADIO SIGNALING SYSTEM, wherein a transmitting system is modulated by means of an electron tube circuit arrangement. The antenna system is continuously supplied with radio frequency energy. An electron tube has its anode and cathode serially arranged with respect to the antenna circuit and its control electrode and cathode included in circuit with a telephone transmitter or other signal controlling device. The amplitude of the radiated energy is thus varied in accordance with the signals.

1,501,831—E. F. W. Alexanderson, filed January 10, 1916, issued July 15, 1924. Assigned to General Electric Company.

RADIO SIGNALING SYSTEM, wherein modulations are produced at the transmitter by a circuit arrangement in which a variable amount of energy is diverted from the transmitting circuit during the production of signals. A circuit is provided in which inductance is neutralized by capacity through resonance. The effective resistance of the energy absorbing circuit is varied in accordance with signals to cause the radiation of suppression of energy.

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1,502,063—W. Schottky, filed November 6, 1920, issued July 22, 1924. Assigned to Siemens and Halske, Aktiengesellschaft, of Siemensstadt, near Berlin, Germany.

ELECTRIC WAVE RECEIVER, wherein a local source is provided for superimposing on the received frequency a local frequency different from that received and arranged to produce a beat frequency current above the limit of audibility. A rectifier is provided which rectifies the beat frequency current. This current is transformed into an alternating current which is then rectified and observed.

1,502,343—B. MacPherson, filed May 25, 1920, issued July 22, 1924. Assigned to Wireless Specialty Apparatus Company, of Massachusetts.

PROCESS AND APPARATUS FOR MAKING ELECTRICAL CONDENSERS of the stack type consisting of alternate layers of mica and foil sheets. A receptacle for the condenser is provided which is substantially higher than the condenser which it is to enclose. The receptacle is filled with melted wax to a point substantially over the top of the condenser. Heat is supplied to the receptacle and to the upper part of the body of wax. The receptacle is then cooled to cause the wax to harden gradually from the bottom upwardly.

1,502,460—G. H. Clark, filed April 26, 1921, issued July 22, 1924. Assigned to Radio Corporation of America, Delaware

RADIO SIGNALING APPARATUS for use with a quenched spark radio transmitter where the quenched spark gap circuit delivers its energy to an intermediate circuit from which it is transferred to the antenna circuit. A mechanical wave changing device is provided for changing the characteristics of each of the circuits simultaneously.

1,502,813—L. Espenschied, filed September 30, 1919, issued July 29, 1924. Assigned to American Telephone and Telegraph Company, of New York.

HIGH FREQUENCY MULTIPLEX SIGNALING SYSTEM, wherein a terminal station is connected by line wire with a distant control station. The antenna circuit is balanced and connected to a conjugate transformer. One of the circuits is tuned to the reception frequency and the other of the circuits to the transmission frequency. A circuit is provided for modulating and demodulating the received and the transmitted currents.

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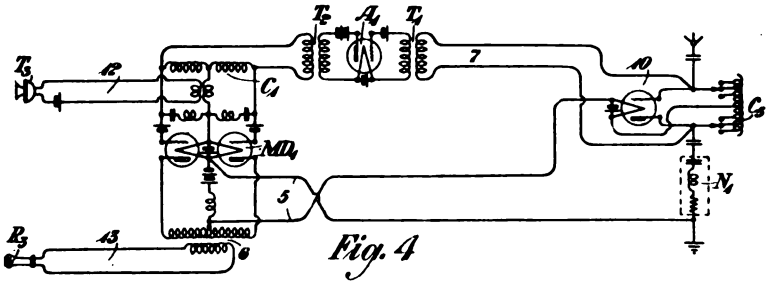

Fig. 4

NUMBER 1,502,813—High Frequency Multiplex Signaling System

1,502,831—S. M. Kintner, filed May 11, 1921, issued July 29, 1924. Assigned to Westinghouse Electric and Manufacturing Company.

SYSTEM OF CONTROL for an arc transmitter, wherein a continuously oscillating circuit is provided coupled with an antenna circuit. The signals are produced by reducing or interrupting the currents in the antenna circuit and simultaneously inserting a sufficient impedance in series with the oscillating circuit to render the antenna circuit ineffective.

1,502,848—F. Conrad, filed July 7, 1920, issued July 29, 1924. Assigned to Westinghouse Electric and Manufacturing Company.

TUNING SYSTEM OF ANTENNAS for radio receiving apparatus where the receiving circuit may be broadly tuned to resonance for a given wave length and next sharply tuned to resonance for increasing the intensity of the received signal. The receiving circuit is provided with a variable inductor and a variable condenser mechanically connected to be simultaneously varied so that the algebraic sum of their reactance remains approximately constant.

1,502,860—D. S. McCrum, filed November 24, 1923, issued July 29, 1924.

VARIABLE CONDENSER where the plates are in the form of cylinders arranged to telescope one within the other forming extended cylindrical capacity areas.

1,502,875—M. I. Pupin and E. H. Armstrong, filed February 10, 1916, issued July 29, 1924. Assigned to Westinghouse Electric and Manufacturing Company.

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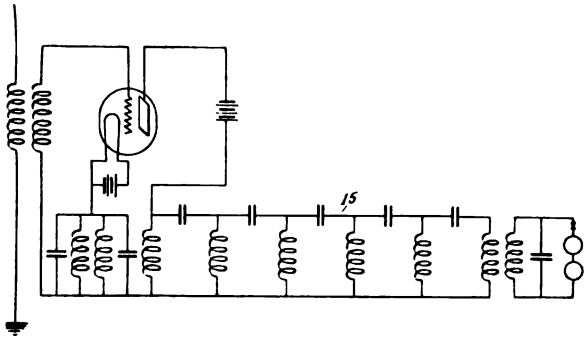

NUMBER 1,502,875—Tone-Producing Radio Receiver

TONE-PRODUCING RADIO RECEIVER, wherein the receiving amplifier is arranged to repeat the incoming waves at an amplitude which varies periodically at an audible frequency so that the resultant electric waves produce a musical note in the receiving system. An electron tube circuit is provided at the receiver with a filter coupling the input and output circuits thereof, the filter including a plurality of like units, each unit comprising two reactances of opposite sign with all the reactances of one sign connected in series and all those of the other sign connected in patallel whereby electrical currents are repeated at a periodically varying amplitude.

1,502,889—H. J. Van Der Bijl, filed January 8, 1918, issued July 29, 1924. Assigned to Western Electric Company.

METHOD OF AND SYSTEM FOR RADIO SIGNALING, by which a large number of messages may be transmitted simultaneously without the use of a correspondingly large number of radio frequency carrier waves. At a receiving station the radio carrier frequency component of the received waves is first eliminated and each of the modulated auxiliary carrier frequency components is transmitted to a modulator from which the various signaling components may be picked out by suitable band filters.

1,503,250—P. E. Sabine, et al., filed August 7, 1920, issued July 29, 1924. Assigned to Paul E. Sabine.

APPARATUS FOR AND METHOD OF RAPID TRANSMISSION OF TELEGRAPHIC MESSAGES by an impulse alphabet, wherein one of two kinds of impulses are transmitted while the other is suppressed. At the receiver the impulse is locally generated and sup plied whereby the receiver is operated in a manner as if both kinds of impulses had been transmitted.

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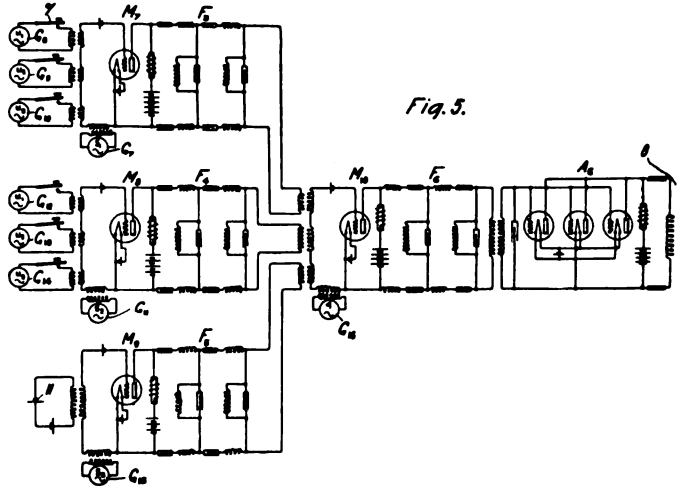

Fig. 5.

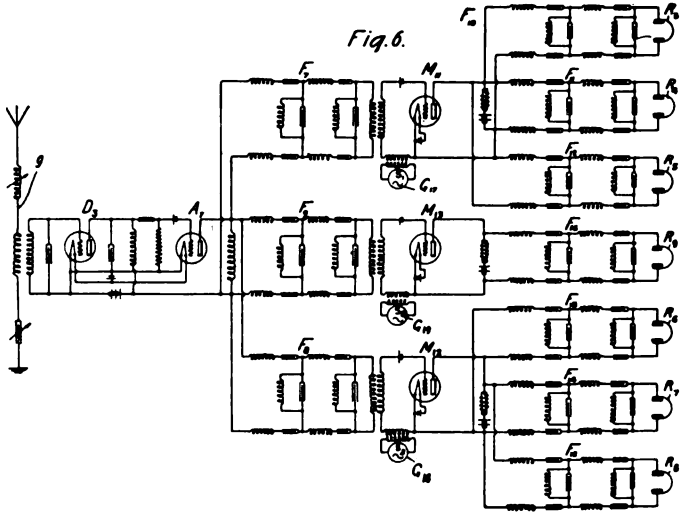

Fig. 6.

NUMBER 1,502,889—Method of and System for Radio Signaling

1,503,308—C. D. Ehret, filed October 22, 1920, issued July 29, 1924.

METHOD OF AND APPARATUS FOR ELECTRICALLY TRANSMITTING INTELLIGENCE in the form of sustained waves without the use of the electrical beat phenomena. The signals are caused to produce a tone frequency and at a point adjacent the production of such tone frequency a magnetically produced sound wave is generated. The sound wave produced by the incoming siganl

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combines with the mechanically-produced sound wave to produce a wave beat of audible frequency for observing the incoming signals.

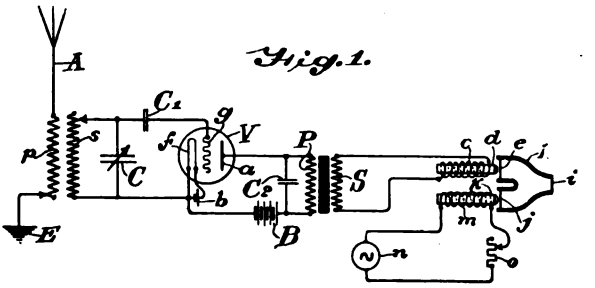

Fig. 1.

NUMBER 1,503,308—Method of and Apparatus for Electrically Transmitting Intelligence

1,503,324—R. Herzog, filed July 12, 1920, issued July 29, 1924. Assigned to C. Lorenz Aktiengesellschaft, of Lorensweg-Berlin, Tempelhof, Germany.

MEANS FOR PRODUCING ELECTRIC OSCILLATIONS BY AN ELECTRIC ARC, whereby uniform burning of the arc is secured by aid of self induction thru a circuit connected in series with the arc. A capacity sufficiently large as not to affect the tuning is inserted in a circuit including the arc and two self induction coils. This circuit does not oscillate independently and the effect of the self inductance upon the antenna is entirely or nearly neutralized, but the circuit operates to stabilize the burning of the arc.

1,503,709—H. M. Pruden, filed April 3, 1923, issued August 5, 1924. Assigned to Western Electric Company, of New York.

VACUUM TUBE CIRCUITS having automatic means for providing for the continuous flow of heating current from a common source thru a plurality of electron tube cathodes when one or more of the cathodes become broken or otherwise removed from the circuit. The invention relates to a bank of electron tubes where the circuit remains operative even tho one of the tubes may be burned out. A relay is provided which substitutes a resistance for the burned-out filament when such filament becomes open.

1,503,765—G. W. Pickard, filed October 8, 1920, issued August 5, 1924. Assigned to Wireless Specialty Apparatus Company, of Massachusetts.

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ELECTRICAL CONDENSER of the stack type where a pair of stacks are arranged transversely of the bottom of the casing with a wedge disposed between the stacks for tightly compressing the stacks against the side walls of the casing.

1,504,002—E. Thomson, filed November 13, 1920, issued August 5, 1924. Assigned to General Electric Company, of New York.

ELECTROSTATIC CONDENSER for high power operation where the condenser is constructed in a stack of thin sheets of alternate conducting material and insulating material. The feature of the invention is the insertion of heat-conducting sheets between the several condenser sections for conveying away heat developed in the condenser.

1,504,303—H. A. Affel, filed June 3, 1921, issued August 12, 1924. Assigned to American Telephone and Telegraph Company, of New York.

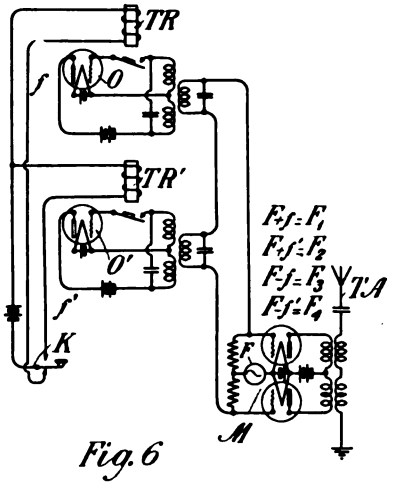

Fig. 6

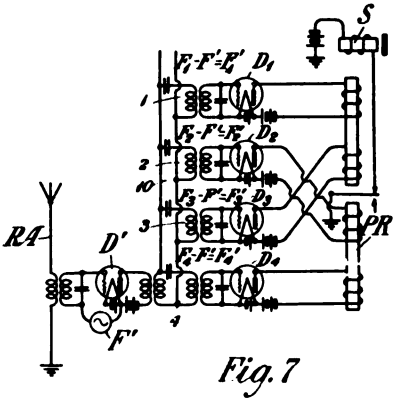

Fig. 7

NUMBER 1,504,303—Method of and Means for Reducing Static Disturbances

METHOD OF AND MEANS FOR REDUCING STATIC DISTURBANCES, wherein the effective transmitting energy is increased in strength by transmitting the energy at several different wave lengths simultaneously, altho the total amount of energy radiated at all of the different frequencies need be no greater than that radiated as heretofore at a single frequency. The conjoint action of the several frequencies is employed to operate a receiving relay for observing the transmitted signal.

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1,504,426—W. M. Bruce, Jr., filed July 7, 1919, issued August 12, 1924.

RADIO TELEGRAPHY, wherein signals are recorded by a variable ink line on a tape. A relay is connected in the output circuit of an electron tube detector with a movable coil connected with a stylus which is moved upon the receipt of signals from a zero position to a marking position by which an ink line is traced upon a moving tape in accordance with the incoming signals.

1,504,462—R. B. Woolverton, filed July 25, 1921, issued August 12, 1924.

TRANSMISSION OF RADIO SIGNALS EMPLOYING CONTINUOUS WAVES, whereby rapid fluctuation of the signal energy is permitted with the consequent production of sharp rapid signals at receiving stations. An arc generator is shown for producing the oscillatory currents which are delivered to a tuned oscillatory circuit and then impressed upon the antenna circuit. The signals are interrupted by destroying the resonance between the antenna circuit and the oscillatory circuit without interrupting the arc.

1,504,537—H. de F. Arnold, filed September 3, 1915, issued August 12, 1924. Assigned to Western Electric Company, Incorporated, of New York.

POWER-LIMITING AMPLIFYING DEVICE for use in a radio receiving apparatus whereby foreign disturbances and heavy static of large magnitude may be reduced to a value not exceeding that of the signaling energy for enabling the signals to be observed thru such interference. The principle of unilateral conductivity is employed by which to distinguish between the signaling energy and heavy static disturbances. In the preferred form of this device the unilateral conductivity is secured by causing part of the circuit to lie in the path of thermionic currents between hot cathodes and cold anodes, said thermionic currents being oppositely directed with respect to said circuit. These thermionic currents are caused to flow by impressing upon their limiting electrodes, in multiple, an electromotive force operating thru a high impedance, said high impedance performing an important function in connection with the power or current-limiting action of the device.

1,504,570—J. O. Mauborgne, et al., filed July 26, 1922, issued August 12, 1924.

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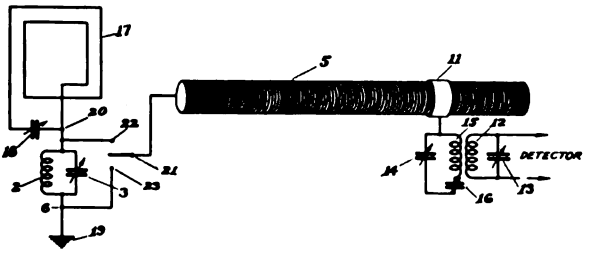

NUMBER 1,504,570—Electrical Signaling

ELECTRICAL SIGNALING, wherein radio signals may be received substantially free of interference by a combination loop antenna and wave coil receiving circuit. The loop antenna is closed thru a variable condenser and connected at one point to ground and to the terminal of a wave coil. The receiving apparatus is coupled by means of a movable ring to the wave coil which is moved along the wave coil to a position for best operation.

1,504,600—Q. A. Brackett, filed January 16, 1919, issued August 12, 1914. Assigned to Westinghouse Electric and Manufacturing Company.

MEANS FOR PROTECTING RADIO OUTFITS FROM STATIC DISTURBANCES, wherein the major portion of the energy of static disturbances is shunted around the receiving apparatus so as to be substantially ineffective in disturbing the receipt of signaling impulses. A pair of rectifying devices are connected in shunt with each other and placed directly across the receiving circuit.

1,504,604—L. W. Chubb, filed May 24, 1921, issued August 12, 1924. Assigned to Westinghouse Electric Company, of Pennsylvania.

SYSTEM OF CONTROL by signaling means of an arc without interrupting the oscillatory condition of the arc. An antenna is employed provided with parallel extended conductors, one set of conductors of which is connected with the arc signaling system, and the other set of conductors to a keying circuit by which the conductors may be included in an absorbing circuit during the production of signals. The sets of conductors in this manner operate as a huge condenser in the absorbing circuit.

1,504,940—G. W. Carpenter, and W. L. Carlson, filed January 29, 1921, issued August 12, 1924.

TELEPHONE RECEIVER CIRCUITS, wherein the telephone head-

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set is electrostatically shielded from radio frequency coupling currents which might stray from other parts of a sensitive electron tube amplifier. The shielded cord as claimed in this patent is a feature of the Navy Type Brandes telephone headset. Howling in sensitive multi-stage amplifiers is avoided by using this type of headset.

FIG. 1

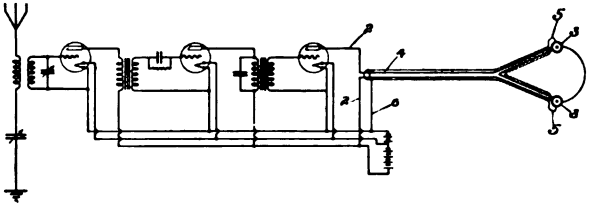
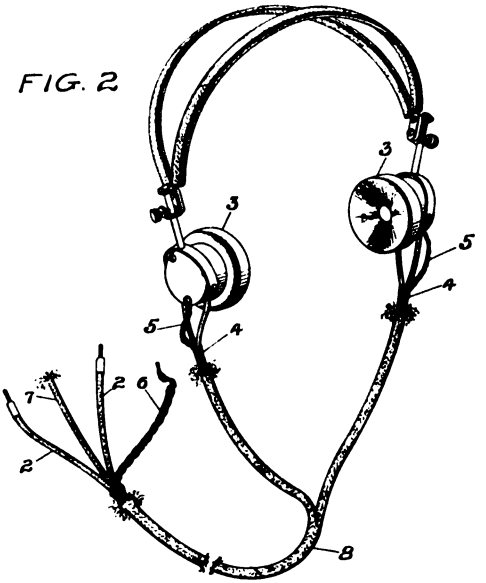


FIG. 2



NUMBER 1,504,940—Telephone Receiver Circuits

1,504,974—C. Reno, filed March 1, 1920, issued August 12, 1924.

SYSTEM FOR TRANSMITTING ENERGY WITHOUT WIRES in a confined path in any direction. A spirally-revolving magnetic field is produced in a pair of symmetrically segmated conductors. A parabolic reflecting circuit is arranged for focusing the magnetic field in a desired direction.

1,505,158—De Loss K. Martin, filed November 23, 1921, issued

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August 19, 1924. Assigned to American Telephone and Telegraph Company, New York.

FREQUENCY CONTROL SYSTEM of carrier waves from a distant radio station. A controlling frequency is transmitted to a receiving station where it is amplified and caused to operate a synchronous motor. The synchronous motor may be employed to operate any desired form of synchronous device.

1,505,600—G. W. Pickard, filed June 8, 1920, issued August 19, 1924. Assigned to Wireless Specialty Apparatus Company, Boston, Massachusetts.

PROCESS AND APPARATUS FOR THE MANUFACTURE OF ELECTRICAL CONDENSERS, where a condenser of the stack type is submerged in a bath of fluid insulating material in a casing and then, while in the bath, a permanent pressure is applied to the stack to drive out the insulating material from between the sheets. The sheets are thus brought into intimate contact with a permanent filler of insulating material around the stack without the stack being exposed to atmosphere after its submerging.

LIST OF RADIO TRADE MARKS PUBLISHED BY PATENT OFFICE PRIOR TO REGISTRATION

(The numbers given are serial numbers of pending applications)

191,177—"VOISOMETER" for radio receiving apparatus. Voisometer Mfg. Co., Mirkwood, Missouri. Claims use since January 23, 1924. Published July 8, 1924.

183,996—"HETEROFLEX" for radio apparatus. Electrical Research Laboratories, Chicago, Illinois. Claims use since November 20, 1923. Published July 8, 1924.

194,212—"POLYTRON" for radio apparatus. Polydyne Corporation, New York City, New York. Claims use since about March 8, 1914. Published July 8, 1924.

194,214—"POLYCOIL" for radio apparatus. Polydyne Corporation, New York City, N. Y. Claims use since about March 8, 1924. Published July 8, 1924.

194,215—"POLYFONE" for radio apparatus. Polydyne Corporation, New York City, N. Y. Claims use since about March 8, 1924. Published July 8, 1924.

194,440—"DYNOFORMER" for radio transformers and coupling

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coils. Electric Service Engineering Corporation, New York, N. Y. Claims use since February 21, 1924. Published July 8, 1924.

194,441—"DYNODON" for radio condensers. Electric Service Engineering Corporation, New York, N. Y. Claims use since February 21, 1924. Published July 8, 1924.

194,464—"MICRO-HELIX" for variable condensers. Packwood Brothers, Lincoln, Nebraska. Claims use since February 15, 1924. Published July 8, 1924.

194,631—"Trade mark consists of the picture of Lafayette, a deceased celebrity," for radio receiving apparatus. Thompson Mfg. Company, Jersey City, New Jersey. Claims use since February 7, 1924. Published July 8, 1924.

166,123—"VOCALION" for radio receiving and transmitting sets The Aeolian Co., New York, N. Y. Claims use since about June 14, 1922. Published July 22, 1924.

170,673—"RESIST-O-METER" for ohmic resistance for radio receiving sets and the like. Scholes Radio and Mfg. Co., New York, N. Y. Claims use since August 14, 1922. Published July 22, 1924.

194,211—"POLY" for radio apparatus and receiving sets. Polydyne Corporation, New York, N. Y. Claims use since about March 8, 1924. Published July 22, 1924.

194,213—"POLYSTAT" for radio apparatus and receiving sets. Polydyne Corporation, New York, N. Y. Claims use since March 8, 1924.

194,785—"MICROFARAD JUNIOR" for condensers. Chelten Electric Co., Philadelphia, Pennsylvania. Claims use since on or about November 13, 1923. Published July 22, 1924.

194,786—"MIDGET VERNIER" for condensers. Chelten Co., Philadelphia, Pennsylvania. Claims use since on or about January 31, 1923. Published July 22, 1924.

194,794—"GR" in ornamental design for radio apparatus and receiving sets. General Radio Co., Cambridge, Massachusetts. Claims use since October 1, 1923. Published July 22, 1924.

194,830—"RADIOLAMP" for loud speakers. Robert E. Wheelan, New York, N. Y. Claims use since March 15, 1924. Published July 22, 1924.

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193,741—"WARE" for radio sets and parts thereof. Wave Radio Corporation, New York, N. Y. Claims use since May 4, 1922. Published July 22, 1924.

193,837—"THE LITTLE KELICOIL" for tuning inductances for radio circuits. Patrick J. Kelly, Philadelphia, Pennsylvania. Claims use since November 11, 1923. Published July 22, 1924.

190,890—"THE CONCERT MODULATOR" for loud speakers. Manhattan Electrical Supply Co., Inc., New York, N. Y. Claims use since August 28, 1923. Published July 22, 1924.

195,331—"VARIO-FARAD" for variable condensers. Thomas E. Binder doing business as Elec-Rad Co., Trenton, New Jersey. Claims use since September, 1923. Published July 22, 1924.

195,718—"PORAT-RADIO" for radio apparatus. Service Radio Corporation, New York, N. Y. Claims use since about April 1, 1924. Published July 22, 1924.

195,755—"MOLODYNE" for radio receiving sets and parts thereof. The Melodyne Co., New York, N. Y. Claims use since April 10, 1924. Published July 22, 1924.

196,258—"GRAND OPERA VOLUMETER" for radio headsets and loud speakers. Radio Industries Corporation, New York, N. Y. Claims use since about December 28, 1923. Published July 22, 1924.

196,466—"HELIOTRON TUBE" for electron tubes. F. S. McCullough Laboratories, Wilkinsburg, Pennsylvania. Claims use since April 26, 1924. Published July 22, 1924.

196,555—"AUDIOTRAN" for radio receiving sets and parts thereof. Scientific Equipment Co., Ridgewood, New Jersey. Claims use since April 17, 1924. Published July 29, 1924.

196,814—"RIT-ONE" for radio apparatus, in ornamental design. Radio Sales Co., New York, N. Y. Claims use since January 1, 1924. Published July 29, 1924.

196,826—"MAGNA-LION" for crystal detectors. Harry B. Schiller, Philadelphia, Pennsylvania. Claims use since May 6. 1924. Published July 29, 1924.

186,703—"THE BULL DOG GRIP" for plugs for radio telephones, in ornamental design. John Firth, New York, N. Y. Claims use since 1921. Published August 5, 1924.

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191,580—"THE ATLANTIC & PACIFIC RADIO CO." in ornamental design, for radio apparatus. Atlantic & Pacific Radio Co. Claims use since June, 1921. Published August 5, 1924.

193,830—"AMPLIOLA" for loud speakers. Alfred Graham & Co., of London, England. Claims use since April 1, 1923. Published August 5, 1924.

193,842—"MANHATTAN" for radio apparatus. Manhattan Electrical Supply Co., New York, N. Y. Claims use since 1889. Published August 5, 1924.

194,532—"SINGLE-TROL" for radio apparatus. G. H. Fischer & Co., Glendale, New York, N. Y. Claims use since March 3, 1924. Published August 5, 1924.

194,891—"UNIDYNE" for radio receiving sets. United Mfg. & Distributing Co., of Chicago, Illinois. Claims use since January 1, 1924. Published August 5, 1924.

195,583—"LINRAD" for radio receiving sets and parts thereof. Lininger Radio Co., Cleveland, Ohio. Claims use since February 11, 1924. Published August 5, 1924.

195,685—"HARCO" for radio receiving sets and parts thereof. Harris & Birdseye, Inc., New York, N. Y. Claims use since October, 1923. Published August 5, 1924.

166,125—"AEOLIAN" for radio transmitting and receiving sets. The Aeolian Co., New York, N. Y. Claims use since June 14, 1922. Published August 12, 1924.

193,527—"LYTTON" for radio set and parts thereof. Walter Lytton, Inc., Chicago, Illinois. Claims use since January 1, 1924. Published August 12, 1924.

193,556—"TRESCO" for radio set and parts thereof. Tri-City Radio Electric Supply Co. Claims use since 1914. Published August 12, 1924.

193,557—"TRESCO" in ornamental design, for radio sets and parts therefore. Tri-City Radio Electric Supply Co. Claims use since 1914. Published August 12, 1924.

194,592—"D" for radio sets and supplies. De Forest Radio Telephone & Telegraph Co. Claims use since 1919. Published August 12, 1924.

194,730—"TRAYDIO" for radio apparatus. Musical Products Distributing Co., Inc. Claims use since March 17, 1924. Published August 12, 1924.

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194,982—"TELEDYNE" for radio receiving sets and parts thereof. Cutting & Washington Radio Corporation. Claims use since February 11, 1924. Published August 12, 1924.

195,118—"ACCURATUNE" for radio apparatus. The Mydar Radio Co., Newark, New Jersey. Claims use since June 15, 1922. Published August 12, 1924.

196,330—"CRYSTALSTAT" for crystal detectors and rectifiers. The Breckenridge Electric Co., Detroit, Michigan. Claims use since March 1, 1924. Published August 12, 1924.

195,981—"MINUTE MAN" for radio sets and parts therefor. Pathe Phonograph & Radio Corp., Brooklyn, New York. Claims use since February 6, 1924. Published August 12, 1924.

195,982—"DEPUTY" for radio set and parts therefor. Pathe Phonograph & Radio Corp., Brooklyn, New York. Claims use since January 30, 1914. Published August 12, 1924.

196,420—"NEUTRADIAL" for radio sets and parts therefor. Broadcast Manufacturers, Inc. Claims use since March 31, 1924. Published August 12, 1924.

196,441—"TELOMONIC" for radio set and parts therefor. Danzer-Jones, Inc., New York, N. Y. Claims use since April 2, 1924. Published August 12, 1924.

196,524—"K-E" for radio apparatus. Kirkman Engineering Corporation. Claims use since February 19, 1924. Published August 12, 1924.

187,998—"WOODEHORN" for radio sets and parts therefor. Inter-Ocean Radio Corporation. Claims use since September 1, 1922. Published August 12, 1924.

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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 12 DECEMBER, 1924 Number 6

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FIELD INTENSITY MEASUREMENTS IN WASHINGTON ON THE RADIO CORPORATION STATIONS AT NEW BRUNSWICK AND TUCKERTON, NEW JERSEY*

By

L. W. Austin

(Physicist, Radio Physical Laboratory, Bureau of Standards, Washington, D. C.)

(Communication from the International Union of Scientific Radio Telegraphy)

Measurements on stations at moderate distances were first taken up with the intention of using their signals as standards for the calibration of the telephone comparator employed in the measurement of European stations. During the preliminary experiments satisfactory constancy was obtained on the signals from WQK (Rocky Point, Long Island), but as warm weather came on, irregularities appeared and the observations were shifted to WII (New Brunswick) since it is nearer and for this reason perhaps less likely to show variations. Observations were taken twice a day at about 10 A. M. and 3 P. M. on a low antenna swung about three meters from the ground between the towers of the main antenna. The radiation height (effective height) is 1.2 meters. The mean of the morning observations for the year, omitting the disturbed days of January, February, and March, gives a field intensity for New Brunswick of 3.06 millivolts per meter. As 10 percent is, under ordinary circumstances the limit of accuracy of these measurements, observations which lie within this limit are considered normal.

In June, 1923, the morning signals from WII, which are more regular than those of the afternoon in summer, on account of the absence of the afternoon absorption, fell below normal on ten days, on three of which the deviation lay between 30 percent and 50 percent. In July, the morning signals were abnormal on seven days, twice being more than 30 percent low, while on July

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17th they rose for an hour or two to 85 percent above normal without apparent cause.

From August 1st to December 30th the signals were more regular. During August they varied in the morning less than 10 percent, except for slight variations on three days. In September they were also below normal on three days, one day being 40 percent low. In October they again exceeded the 10 percent variation slightly on three days. In November the three variations were all high, the greatest deviation being 24 percent. In December there was one low variation of 21 percent and one high, 63 percent.

Regular observations were begun on WGG (Tuckerton), in October. This station averages about 12 percent below WII in intensity, giving 2.7 millivolts per meter. From the tables it will be seen that the variations are of the same order as those of New Brunswick. Figure 1 shows the monthly averages of the New Brunswick (WII) observations. Figure 2 shows the daily record for WII in July, the crosses being the 10 A. M. observations and the circles those at 3 P. M. The tables give the complete daily observations for both stations.

During December there was no very cold weather in Washington, but with the cold waves of January there was a marked

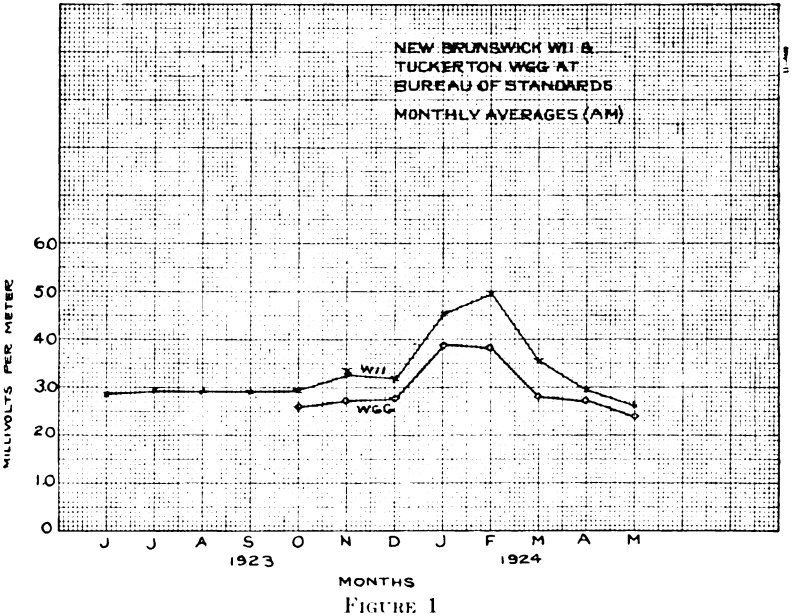

FIGURE 1

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increase in the irregularities of WII and WGG and in a lesser degree of WQK. This last station was only observed occasionally.

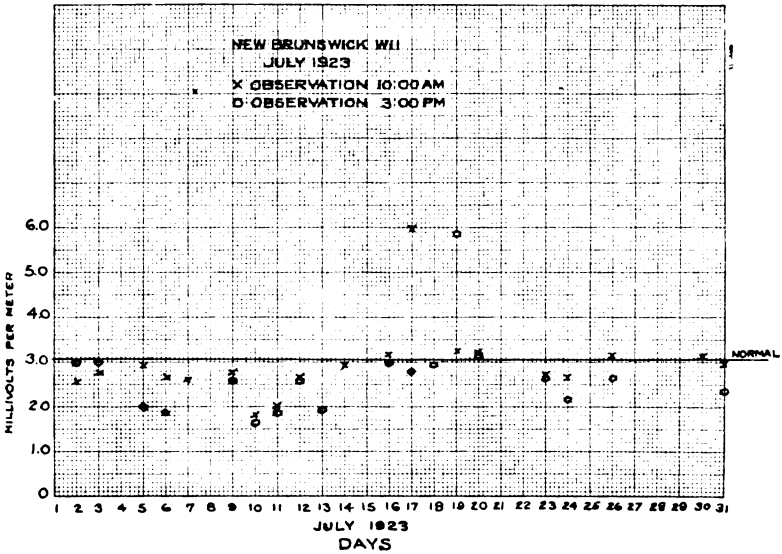

FIGURE 2

At the time of the passage of each of the cold waves there was an increase in signal strength, especially during the severe cold toward the end of the month. In order to show this more clearly the maximum signal of WII or WGG at the two observation times and the reciprocal of the temperature are plotted in Figure 3. The observations on the signals and the 8 o'clock A. M. temperatures are given in the January Table.

During the moderate cold of January 7th and 15th, sometimes only one of the two stations seemed to be affected and the high readings were observed for only a part of the time, but during the extreme cold of January 22d and 27th both of the stations regularly observed as well as WQK and WSO (Marion, Mass.) remained high thruout the day, and in the last case for several days. There were also at these times marked shifts in the apparent direction of the stations even in the forenoon when their bearings are usually correct. NSS (Annapolis) 53 km. away showed no increase in signal strength during these times. The end of the cold weather of January did not at once restore receiving conditions to the uniformity of the autumn and early winter, but left an instability which persisted thru the comparatively mild cold of February and a portion of March. During this time

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the signals from WII and WGG fluctuated frequently, going to more than twice their normal strength for a few hours and then returning to normal without any obvious relation to weather conditions.

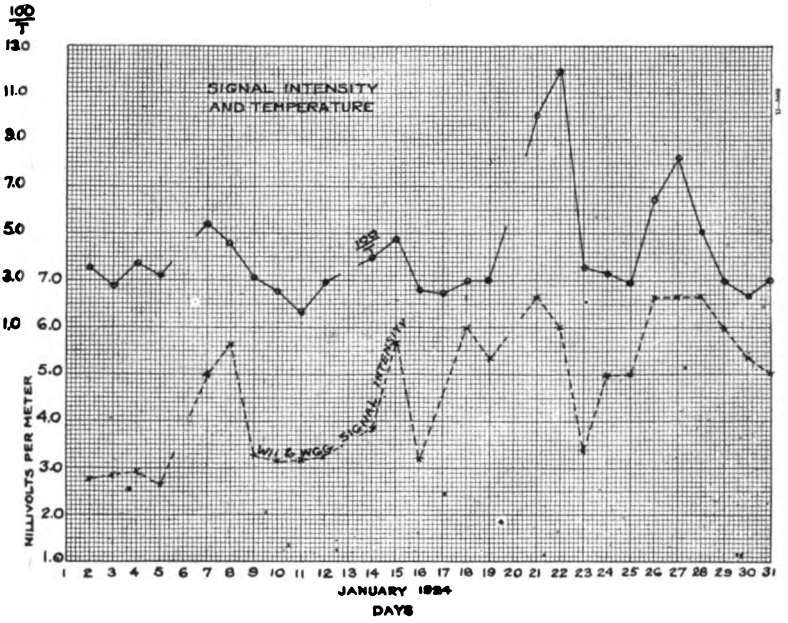

FIGURE 3

Since March 19th these large irregularities have ceased and transmission seems to have returned to a stable condition with nearly the same average signal strength as in the autumn. The observations which have been taken on Annapolis and on the Radio Corporation stations at various distances indicate that large variations in intensity do not generally occur at 50 km., that they are large at distances of 250 to 300 km. and again become less at 400 to 700 km.

The few night observations which have been taken show somewhat higher and less regular values than those taken during the day.

It is too early to give any definite explanation of the variations observed. Their cause is evidently atmospheric and the connection with the cold waves of January suggests either that the part of the atmosphere concerned in the variations lies much below the Heaviside surface (80 or 100 km.) or that weather phenomena are in some way correlated with atmospheric action at much greater heights than has been hitherto supposed.

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FIELD INTENSITY OF NEW BRUNSWICK WII IN MILLIVOLTS PER METER

Date	1923					
	June		July		August	
	A.M.	P.M.	A.M.	P.M.	A.M.	P.M.
1	2.86	....	....	....	3.13	3.13
2	....	....	2.53	2.99	3.13	2.53
3	....	3.26	2.79	2.66	2.39	....
4	2.59	2.86	....	....	3.13	....
5	....	....	2.93	1.99	....	....
6	....	....	2.66	1.86	....	1.93
7	2.86	2.79	2.59	....	3.08	....
8	1.73	1.94	....	....	....	....
9	1.46	2.66	2.79	2.59	....	....
10	....	....	1.79	1.66	....	....
11	2.39	2.46	1.99	1.86	2.99	....
12	....	2.66	2.66	2.66	....	....
13	2.66	2.19	....	1.93	2.79	2.99
14	2.53	2.66	2.93	....	....	1.99
15	2.79	1.94	....	....	2.99	2.13
16	1.93	....	3.19	2.33	....	2.33
17	....	....	6.00	2.79	2.39	1.93
18	2.53	2.16	....	2.93	2.66	....
19	2.59	3.06	3.26	5.88	....	....
20	2.93	....	3.13	3.13	3.06	3.13
21	2.66	2.33	....	....	....	1.66
22	2.59	2.86	....	....	2.99	2.86
23	2.79	....	2.66	2.66	2.53	2.86
24	....	....	2.66	2.16	....	2.79
25	2.93	2.66	....	....	2.99	....
26	3.06	2.26	3.13	2.66	....	....
27	....	2.33	....	....	3.13	2.99
28	2.99	2.66	....	....	2.99	3.13
29	2.86	2.35	....	....	3.13	2.99
30	2.66	....	3.13	....	2.99	2.86
31	....	....	2.99	2.33	....	2.46
Average	2.86	2.53	2.94	2.62	2.91	2.60

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Field Intensity of New Brunswick WII in Millivolts per Meter

1923

Date	Sept.		Oct.		Nov.	
	A.M.	P.M.	A.M.	P.M.	A.M.	P.M.
1	3.26	....	3.13	....	3.13	2.99
2	....	....	3.19	3.13	3.19	2.59
3	....	....	3.19	3.13	3.26	2.99
4	....	....	....	3.06	....	....
5	3.06	....	3.13	2.33	....	2.86
6	2.66	2.93	2.99	3.13	2.99	3.06
7	....	2.99	....	....	3.13	3.26
8	2.99	....	3.13	2.99	3.26	3.19
9	....	....	3.13	2.99	3.33	3.33
10	3.06	2.59	....	2.99	3.33	3.06
11	....	3.06	2.99	2.59	....	....
12	3.13	2.79	2.66	....	....	....
13	....	....	....	2.66	....	....
14	2.53	3.19	....	....	3.26	3.26
15	....	....	2.99	....	3.33	3.33
16	....	....	....	....	3.19	3.66
17	1.86	3.13	....	....	3.33	3.66
18	2.59	2.86	....	2.13	....	....
19	2.99	3.29	2.26	2.33	3.53	3.53
20	....	2.53	2.59	2.66	3.46	3.46
21	3.26	2.93	....	....	....	3.73
22	3.13	3.13	2.79	2.26	3.79	2.66
23	....	....	2.59	3.19	....	2.53
24	3.26	2.59	....	....	3.26	2.59
25	3.19	3.29	....	....	....	....
26	....	2.59	3.19	2.99	3.19	....
27	....	....	2.99	3.19	2.99	2.99
28	....	3.19	....	....	3.19	3.19
29	....	2.99	2.79	3.19	....	....
30	....	....	3.13	3.13	2.99	2.99
31	....	....	....	3.06	....	....
Average	2.91	2.96	2.94	2.84	3.26	3.13

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Field Intensity of New Brunswick WII in Millivolts per Meter

1923-1924

Date	Dec., 1923		Jan., 1924			Feb., 1924	
	A.M.	P.M.	A.M.	Temp. ° F. at 8 A.M.	P.M.	A.M.	P.M.
1	....	3.19	....	....	....	6.65	3.26
2	....	....	2.66	29°	2.33	5.66	5.66
3	....	....	3.06	36	2.66	....	....
4	....	....	....	27	3.13	....	6.00
5	....	3.26	....	31	3.26	5.00	3.99
6	....	3.13	....	..	....	3.66	....
7	....	3.19	5.00	19	3.13	4.33	6.33
8	5.00	3.26	....	22	3.06	6.00	5.66
9	....	....	3.25	33	....	6.33	5.66
10	....	2.33	3.13	37	3.13	....	....
11	3.13	2.13	....	63	3.19	....	6.65
12	....	2.66	3.26	35	....	5.33	4.33
13	2.99	2.99	....	..	....	3.86	3.33
14	....	2.99	3.66	26	3.06	3.73	3.66
15	2.26	2.66	....	21	....	6.33	5.00
16	....	....	3.13	40	....	5.00	5.33
17	....	2.59	4.66	41	3.33	....	....
18	2.53	....	5.66	35	6.00	....	5.00
19	....	....	5.32	33	3.26	....	6.65
20	....	3.19	....	..	....	3.66	5.33
21	....	3.13	....	10	6.65	5.00	5.00
22	....	3.19	6.00	8	4.33	....	....
23	....	....	3.26	29	3.26	....	6.00
24	3.19	....	5.00	30	5.00	....	....
25	....	....	5.00	35	....	3.19	5.00
26	3.13	3.06	5.66	16	5.00	3.19	3.19
27	....	2.99	6.65+	12	....	6.65+	6.65+
28	....	3.13	6.65+	20	4.66	5.53	6.65+
29	....	3.66	5.32	34	6.00	5.00	....
30	....	....	3.26	42	5.33	....	....
31	....	3.19	....	33	4.66	....	....
Average	3.18	3.00	4.52		4.02	4.94	5.19

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Field Intensity of New Brunswick WII in Millivolts per Meter

1924

Date	March		April		May	
	A.M.	P.M.	A.M.	P.M.	A.M.	P.M.
1	....	3.20	....	3.33	3.20	1.66
2	....	....	3.19	3.19	3.00	2.86
3	4.00	4.33	....	2.53	2.33	1.68
4	....	3.20	....	3.20	....	....
5	....	3.26	3.19	3.20	3.06	2.33
6	3.53	3.26	....	....	3.26	3.13
7	3.26	3.26	....	4.66	....	3.13
8	....	3.26	....	3.13	....	1.40
9	....	....	4.00	3.26	2.33	2.66
10	3.33	3.13	....	3.26	2.33	2.53
11	5.00	3.66	....	3.20	....	....
12	3.20	3.26	....	2.86	....	2.33
13	3.93	3.26	....	....	2.33	2.33
14	3.20	3.86	2.53	3.00	2.00	1.88
15	5.00	5.00	3.20	2.86	1.88	3.00
16	....	....	....	2.53	2.00	2.00
17	....	4.33	....	3.20	2.33	3.20
18	4.66	5.00	....	3.20	....	....
19	....	4.06	....	3.20	3.19	3.06
20	....	3.26	....	....	3.26	3.19
21	....	3.20	....	3.26	2.00	3.06
22	3.26	3.26	....	3.20	3.20	3.20
23	....	....	....	2.66	3.06	3.06
24	3.20	3.20	3.00	3.06	2.26	....
25	....	3.33	3.13	2.26	....	....
26	2.66	3.20	3.26	5.33	3.00	2.50
27	....	3.20	....	....	2.30	2.30
28	....	3.20	3.20	3.00	....	2.30
29	2.53	3.20	1.66	2.66	2.30	2.50
30	....	....	2.00	1.66	....	....
31	2.66	2.53	....	....	2.50	2.60
Average	3.56	3.51	2.95	3.11	2.60	2.55

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Field Intensity of Tuckerton WGG in Millivolts per Meter

1923

Date	October		November		December	
	A.M.	P.M.	A.M.	P.M.	A.M.	P.M.
1	....	....	2.63	2.30	....	2.43
2	....	....	2.63	2.63	....	....
3	....	....	2.63	2.50	....	....
4	....	....	....	....	....	....
5	....	....	....	....	2.76	2.63
6	....	2.50	2.63	....	2.63	2.50
7	....	....	....	2.50	....	3.16
8	2.56	2.50	....	2.43	2.96	....
9	....	2.43	2.63	2.76	....	....
10	2.50	2.56	2.63	2.43	....	....
11	....	....	....	....	....	2.30
12	2.56	....	2.56	2.63	3.36	3.36
13	....	2.50	2.63	2.50	....	3.30
14	....	....	2.63	2.56	2.70	2.50
15	....	....	2.56	2.63	2.96	3.01
16	2.56	2.50	2.56	3.09	....	....
17	....	2.50	2.70	2.63	2.90	2.76
18	....	2.43	....	....	2.16	....
19	2.16	2.23	2.90	2.76	....	....
20	....	2.50	2.83	2.63	....	2.63
21	....	....	....	2.83	....	2.90
22	2.50	2.56	3.16	2.56	....	3.16
23	2.56	....	....	2.30	....	....
24	2.50	2.30	2.96	2.56	....	....
25	2.96	2.63	....	....	....	....
26	2.83	3.00	3.09	....	....	....
27	2.76	2.96	2.76	2.63	2.63	2.63
28	....	....	....	2.63	4.93	....
29	2.56	2.76	....	2.96	2.96	4.60
30	2.63	2.76	....	2.56	....	....
31	2.70	2.50	....	....	2.63	2.63
Average	2.60	2.56	2.72	2.60	2.77	2.79

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Field Intensity of Tuckerton WGG in Millivolts per Meter

1924

Date	January		February		March	
	A.M.	P.M.	A.M.	P.M.	A.M.	P.M.
1	....	....	5.56	2.96	2.76	2.96
2	3.43	2.56	4.26	3.16	....	....
3	2.70	2.56	....	....	3.54	3.30
4	....	....	2.96	4.93	2.96	....
5	2.63	....	4.26	3.60	2.76	....
6	....	....	2.96	....	2.70	2.76
7	....	....	3.16	4.93	....	2.76
8	5.66	3.16	4.93	....	....	2.63
9	....	....	5.24	4.93	....	....
10	....	....	....	....	2.90	2.63
11	3.16	....	3.20	5.56	2.76	3.16
12	....	2.96	4.26	3.30	....	2.63
13	....	....	4.80	....	3.20	2.96
14	....	3.80	2.76	3.60	2.63	3.36
15	2.96	5.56	4.60	3.60	3.80	2.30
16	....	2.96	4.26	4.26	....	....
17	....	3.73	....	....	2.70	3.30
18	....	4.93	3.30	5.90	....	2.96
19	3.93	2.96	....	5.56	2.83	3.20
20	....	....	1.20	....	2.70	2.76
21	2.76	3.93	2.90	3.94	2.50	2.63
22	5.56	....	....	....	2.96	2.90
23	2.96	2.96	4.93	5.90	....	....
24	4.60	4.93	....	....	2.83	2.83
25	4.26	4.93	2.96	....	2.90	2.90
26	4.93	4.26	....	....	2.23	2.70
27	....	....	....	6.56	2.23	2.90
28	4.93	....	....	5.56	3.00	2.63
29	2.96	4.60	....	4.26	2.50	2.63
30	3.60	4.93	....	....	....	....
31	4.93	3.94	....	....	2.43	2.30
Average	3.88	3.88	3.82	4.58	2.81	2.84

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Field Intensity of Tuckerton WGG in Millivolts per Meter

1924

Date	April		May	
	A.M.	P.M.	A.M.	P.M.
1	2.83	2.83	2.96	1.78
2	2.76	3.06	2.53	4.26
3	2.83	2.03	1.98	1.80
4	3.16	....	....	....
5	2.76	3.93	2.83	2.30
6	....	....	3.06	2.30
7	2.96	....	2.76	2.76
8	2.30	....	1.66	2.30
9	4.26	3.06	2.30	2.63
10	2.76	2.96	1.66	2.50
11	2.96	2.70	....	....
12	2.83	2.50	1.66	....
13	....	....	2.30	2.30
14	2.30	....	....	2.53
15	2.76	2.63	2.50	2.96
16	2.76	2.50	2.30	2.30
17	2.96	3.16	2.30	3.16
18	2.76	2.76	....	....
19	2.30	3.16	3.06	3.03
20	....	....	2.70	2.75
21	2.63	2.90	1.98	2.75
22	2.76	....	2.99	3.30
23	2.83	2.96	3.06	3.60
24	....	2.63	2.00	....
25	2.76	2.96	....	....
26	....	5.23	2.70	2.76
27	....	....	2.50	2.30
28	2.83	2.76	1.66	2.60
29	1.70	2.76	2.30	2.10
30	1.83	1.65	....	....
31	....	....	....	2.60
Average	2.73	2.91	2.40	2.65

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TRANSMISSION DATA

	Frequency kc.	Approx. Antenna Current amp.	Distance from Washington km.	Radiation height m.
New Brunswick (WII)	22.04	600	281	66.0
Tuckerton (WGG)	18.85	470	251	67.5

Bureau of Standards
Washington, D. C.

SUMMARY: The paper gives the results of field intensity measurements at the Bureau of Standards on the transatlantic radio stations at New Brunswick and Tuckerton, New Jersey. The summer measurements are somewhat weaker than those of winter and generally the afternoon signals are slightly weaker than those of the forenoon. The most interesting fact observed was the great increase in signal strength which accompanied the passage of the cold waves of January, 1924. This, if verified, will be of special interest as showing a definite connection between radio transmission and the usual weather phenomena.

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THE HIGH POWER STATION AT MALABAR, JAVA*

COMPILED BY
THE EDITOR FROM PHOTOGRAPHS AND NOTES RECEIVED FROM
DR. CORNELIS J. DE GROOT
(DIRECTOR OF RADIO, DUTCH EAST INDIES, BANDOENG, JAVA)

It has been deemed desirable to present to the readers of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, in the form of a permanent record, a description of the erection, equipment, and operation of what is to-day the highest power transmitting station in the world, and one which required for its successful construction an unusual amount of perseverance and ingenuity. Situated at Malabar, in the mountains of Java, and at a considerable distance from all cities, the problems of transportation were extreme. This point will be particularly appreciated when the construction of the gigantic arc and its massive magnetic circuits is considered below. It should also be noted that a good part of this construction was necessarily carried out either during wartime or at a period when the disorganization resulting from the war greatly handicapped the engineer-constructors.

The object of the station was to connect the principal cities of Java by direct radio communication with a correspondent station at Kootwijk, Holland, at a distance of twelve thousand kilometers (or seven thousand five hundred miles). It was desired to establish a twenty-four hour per day commercial service over this hitherto unattempted distance, and with reception in a tropical locality where atmospheric disturbances were known to be unusually heavy.

The illustrations of this compilation are arranged in a generally chronological order, thus giving a clear idea of the sequence of events during the construction and an indication of the procedure followed in the erection of large stations of this type.

The station is located at one end of a steep ravine in the mountains, in what was originally a wilderness. The mountains at each side of the ravine serve as antenna supports, as will be

*Received by the Editor, December 1, 1923.

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FIGURE 1

described in further detail below. It became necessary, therefore, to build the village shown in Figure 1 in the foreground before work could be begun on the station building shown in the rear. The character of the country is also clearly depicted in this figure. A closer view of the station building in process of erection is given in Figure 2. A temporary roof is in place, and

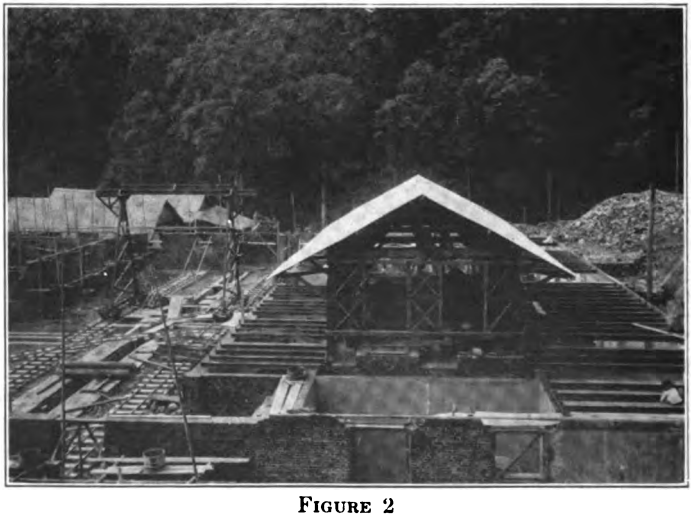

FIGURE 2

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a crane for the erection of the large arc is in place, as are the generator foundations.

Unusual precautions were taken to ensure continuity of power supply for this station. Two double 25,000-volt overhead lines carry power to the station, one from the north and the other from the south. Each of these connects both to steam-driven plants and water power plants. Thus, even in the dry season, complete reliability is obtained by dependence on the steam-driven plants. A sub-station is provided at the end of these lines, containing four transformers, each of 2,000 kilovolt-amperes, and for reducing the voltage from 25,000 to 6,000. One of these transformers is a spare. Appropriate oil switch gear is also provided in the sub-station. The layout of the horn gap arresters in this sub-station is shown in Figure 3, and the core and windings of a 2,000 kva. transformer in Figure 4. In a fireproof cellar within the transmitting station are mounted the three 250-kva. 6,000-

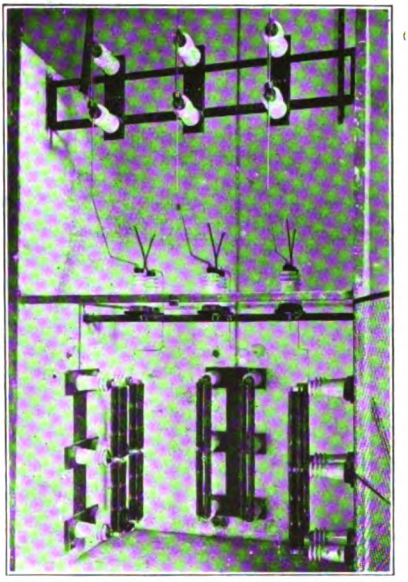

FIGURE 3—Horn Gap Arresters.

220-volt transformers shown in Figure 5, which transformers feed the motor generators in the station up to 100 kw. The large motor generators are operated directly on the 6,000-volt circuits.

There are two hydraulic power plants near the station, each of 200 kilowatts output, which act as a reserve source of power. They also form a spare source for supplying motor power to the

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FIGURE 4—Core and Windings of 2,000 KVA Transformer

FIGURE 5—6,000-220 volt Transformers in Fireproof Cellars for Small Motor Generators

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water pumps for use in case of fire, and can supply the radio station in case the overhead lines fail or are cut. On the power available from these plants, it is possible to reach Holland at night and all of India, Cavite, or Japan by day or night. The water for these plants comes from a small river flowing down the ravine which is led into settling pools for the removal of sediment. An old wooden conduit which conducts the water from the settling ponds is illustrated in the back of Figure 6, while the

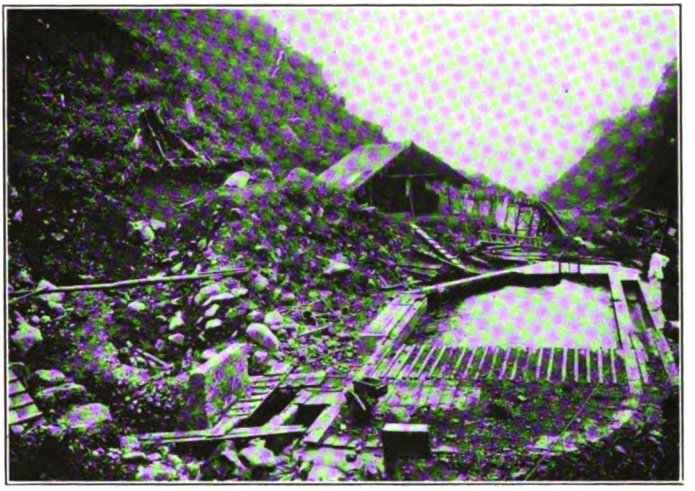

FIGURE 6

beginning of the new tunnel is shown in the left-hand front of the same figure. This same tunnel ends in an open conduit which leads into the first reservoir, and appears in Figure 7. In Figure 8 is seen the process of filling the reservoir, from which the wooden pipe shown in Figure 9 leads down to the first turbine. This wooden pipe— a war necessity—has stood up well under the pressure. It was obtained from Australia during wartime, steel pipe being unobtainable. The hydraulic plant itself is shown in Figure 10. The water wheel, at the left, was taken from a 20-year-old gold placer mine installation in Sumatra during war time. New regulators for speed control have been recently added. This wheel drives a 1,200-volt 150-kilowatt direct current generator obtained in Japan, and the only type available during the war. This generator is now connected by cable to the new radio station, where it can feed the smaller arcs. It is planned shortly to change this generator over into a 6,000-volt three-phase ma-

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chine, so that the entire power supply to the radio station will be of uniform character.

After leaving the first hydraulic plant, the water flows to the second turbine shown in Figure 11. This turbine runs two

FIGURE 7—End of Tunnel with Open Conduit to First Reservoir

FIGURE 8—Filling Reservoir

2,000-volt three-phase alternators which are also to be replaced by a 6,000-volt three-phase alternator. The current from these alternators passes thru an overhead line to the fire pump station, which is therefore provided with an independent source of power for emergencies. This fire pump is placed near the new radio station, and from this point the power is run by

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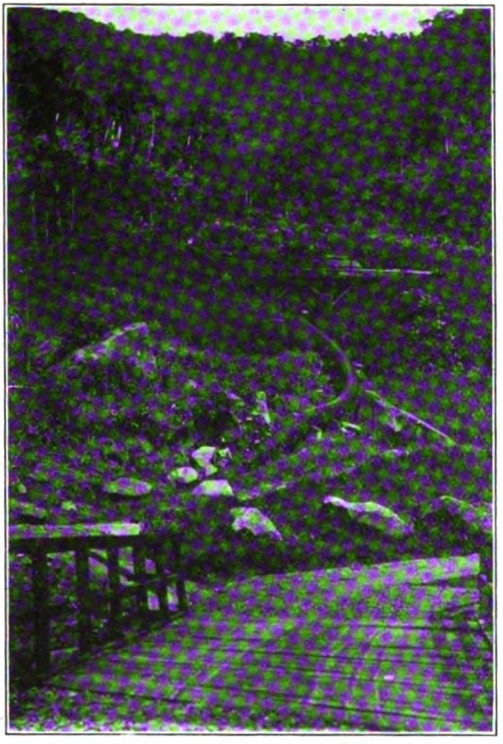

FIGURE 9

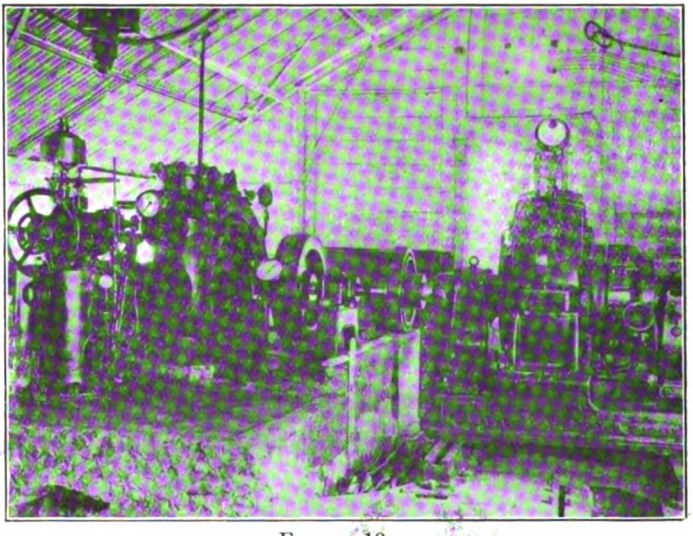

FIGURE 10

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cable to the station where some motor-generators for the small arcs are fed by it. These are motor generators of 200 kw., 150 kw., and 100 kw. direct current output respectively. The first two were purchased in Japan, whereas the latter was obtained from America in wartime. In Figure 12 is illustrated a 6,000-volt 200- kilowatt generator at the radio station. Work was started on a 100-kilowatt Federal Telegraph Company arc. By 1918, this arc had reached the antipodes, and by 1919 it had reached Holland regularly at night time. Its use with Holland was discontinued in 1922 after completion of the 400 kw. Telefunken alternator.

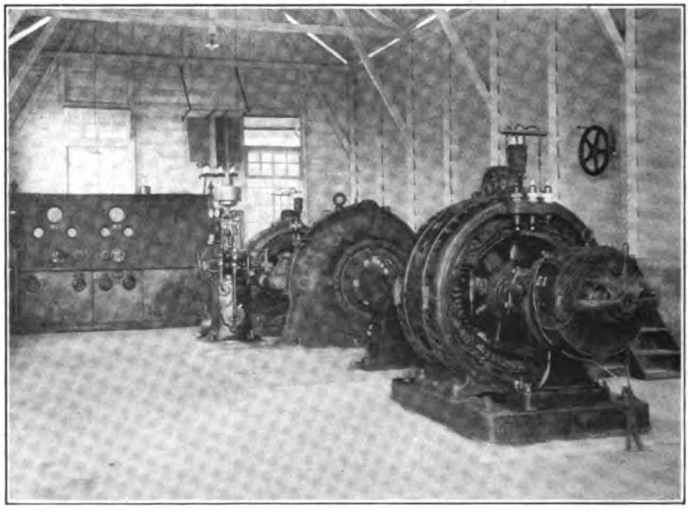

FIGURE 11

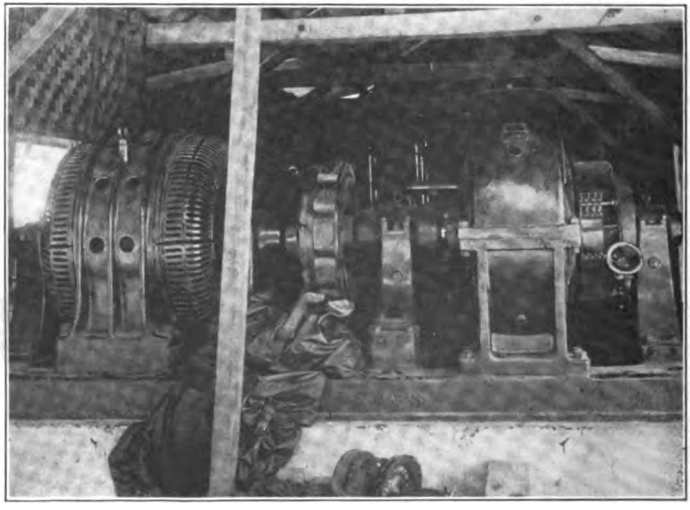

FIGURE 12—6,000-volt Motor Generator, 200 KW.

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The machine room floor on which most of the preceding machines are now placed is shown in Figure 13. The switchboards to the right, for control purposes, were fabricated at the shops of the Malabar station. For supplying power to the high power arc at the new radio station, there are provided two large motor generators, shown in Figure 14. Each of these is capable of delivering 1,200 kilowatts at 3,500 volts direct current. They are

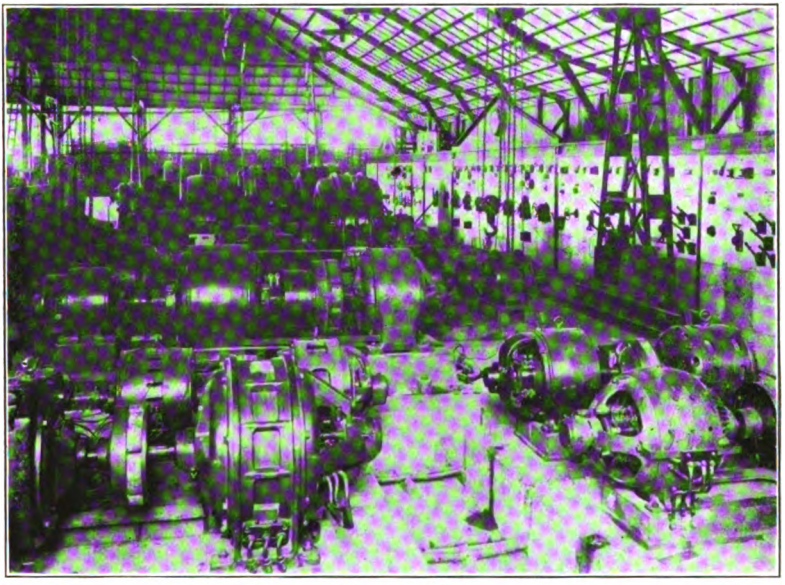

FIGURE 13

now worked in parallel every day to supply a power of 2,400 kilowatts to the arcs whereby 1,200 kilowatts are fed into the antenna. Two more of the motor-generators are to be added next year, one being a spare. It will thus become possible to supply 3,200 kilowatts to the arc, obtaining about 1,600 kilowatts in the antenna. These motor generators were supplied by the General Electric Company and operate very satisfactorily. The quick-break switches for these motor generators are mounted in the cellar, and are shown in Figure 15. These also work well and prevent flash-over on an occasional short circuit on the generators resulting from the normal operations of the arcs without any series resistance.

The lower arc core was erected, using a hydraulic jack. Its cross section is of logarithmic shape so as to produce a maximum flux. The huge magnetic circuit of the arc is built up of numerous

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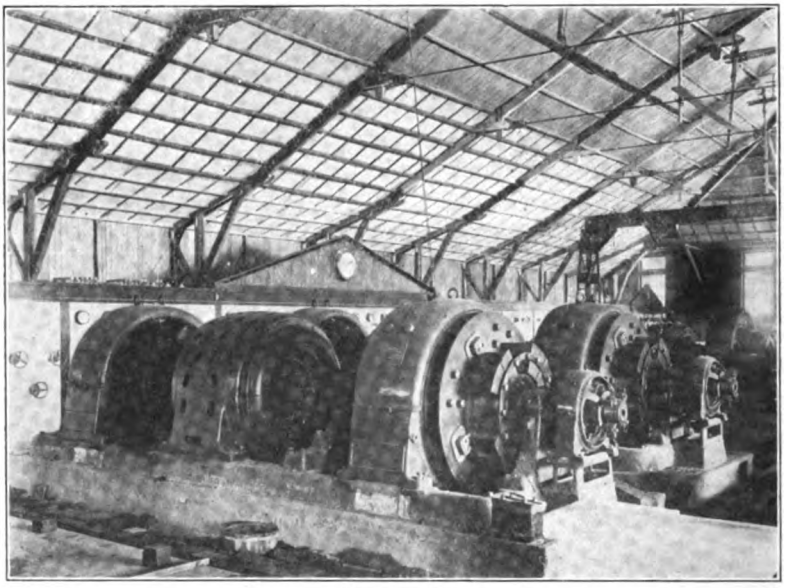

FIGURE 14

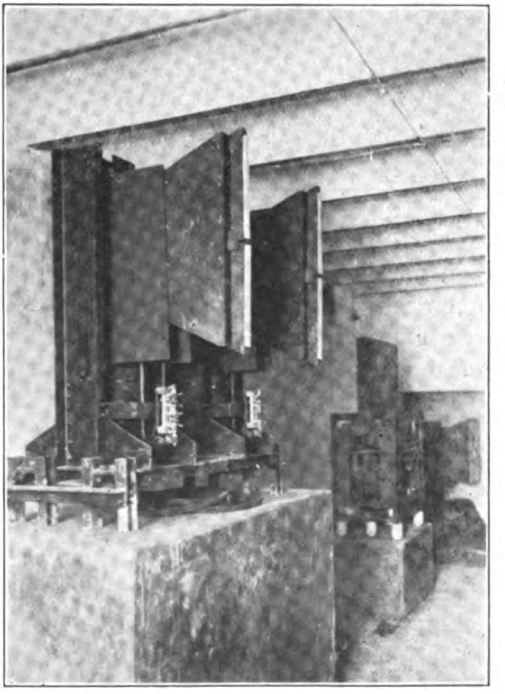

FIGURE 15—Quick Break Switches for Motor Generator

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large pieces of cast iron, each weighing between 1,000 and 6,000 kilograms (2,200 and 13,200 pounds), all held together by steel plates and bolts. This construction was required because of the limitation on the size of castings which could be produced and machined in the Javanese foundries. The method of assembling the magnet core is clearly indicated in Figure 16, which represents the

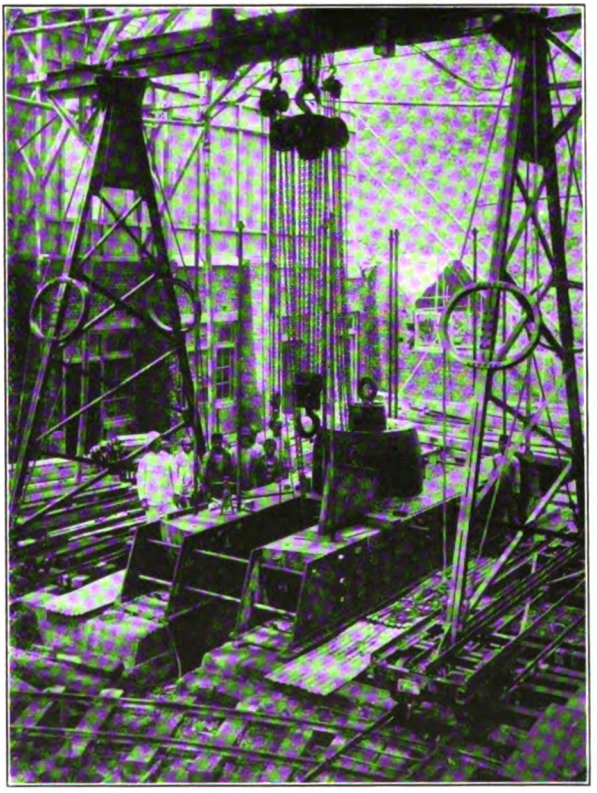

FIGURE 16—Method of Assembling Magnet Core. Lower Core Complete Except Conical Tip and Base.

lower core complete except for the conical tip and the base. The core can be moved vertically by hydraulic pressure and weighs 25,000 kilograms (nearly 28 tons). The lower magnet coil, which also acts as a choke coil in the arc supply circuit, is placed in an oil tank illustrated in Figure 17. After this tank was placed in position and portions of the vertical iron pillars forming the magnetic circuit were in place, the photograph of Figure 18 was taken, the building having progressed considerably by that time.

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FIGURE 17—Oil Tank for Lower Magnet Coil

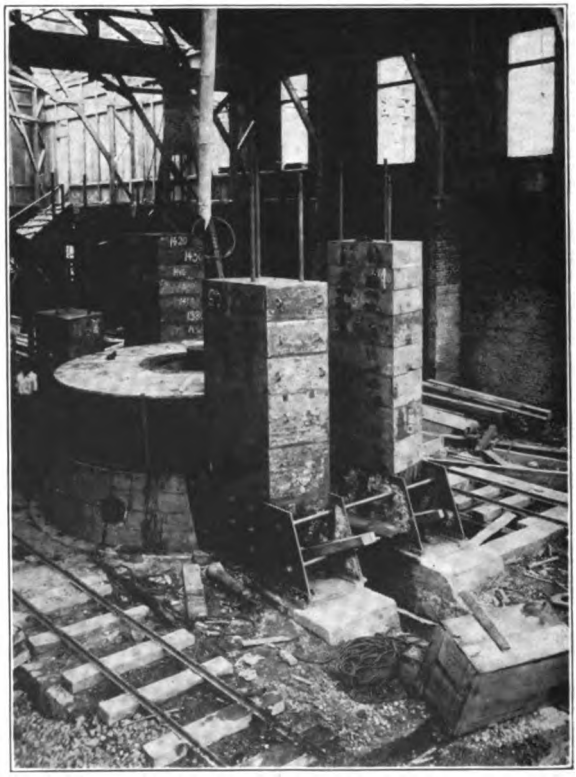

FIGURE 18—Oil Tank and Vertical Magnet Pillars in Place around Lower Coil

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The construction of the lower magnet coil was a difficult task. The beginning of the winding is shown in Figure 19, and Figure 20 shows the winding nearly completed. The actual conductor was made up of nine square-cross-section copper conductors, 7 mm. by 7 mm. (0.3 by 0.3 inch), these being com-

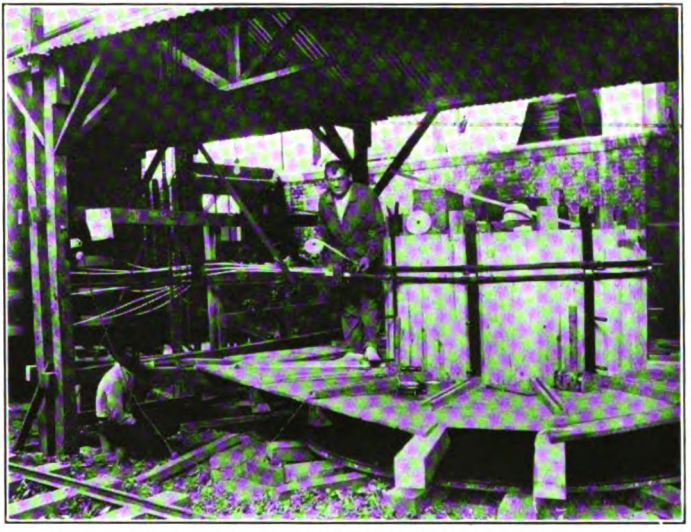

FIGURE 19—Beginning Lower Magnet Coil

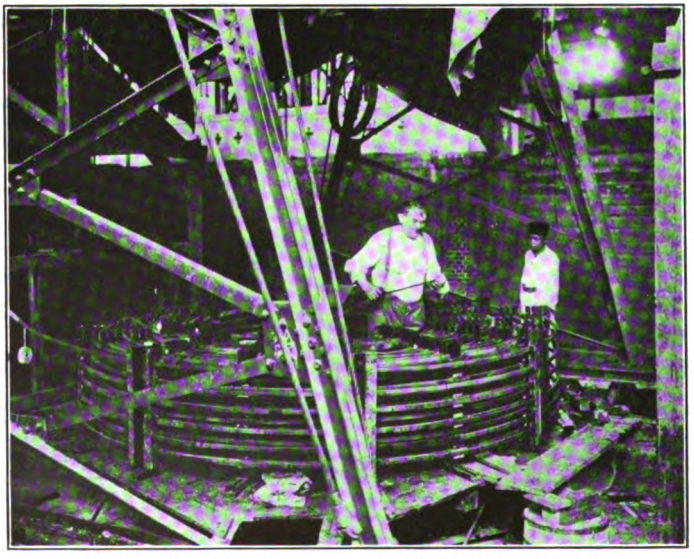

FIGURE 20—Lower Magnet Coil Nearly Completed

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bined into a single conductor of 450 square mm. (1.4 square inch) cross section. It was insulated to withstand 60,000 volts from the arc. During a year of rough use, no trouble has developed in this coil. The copper in this coil weighs 10,000 kilograms (22,600 pounds). The coil is wound in seven concentric cylindrical layers for easy erection and repair.

The next group of photographs shows serially the growth of the large arc. In Figure 21, the lower coil is in place, together with the upper magnet core, and a part of the upper bridge. The four-pillar construction of the outside elements of the magnetic circuit is clearly shown in Figure 22. This type

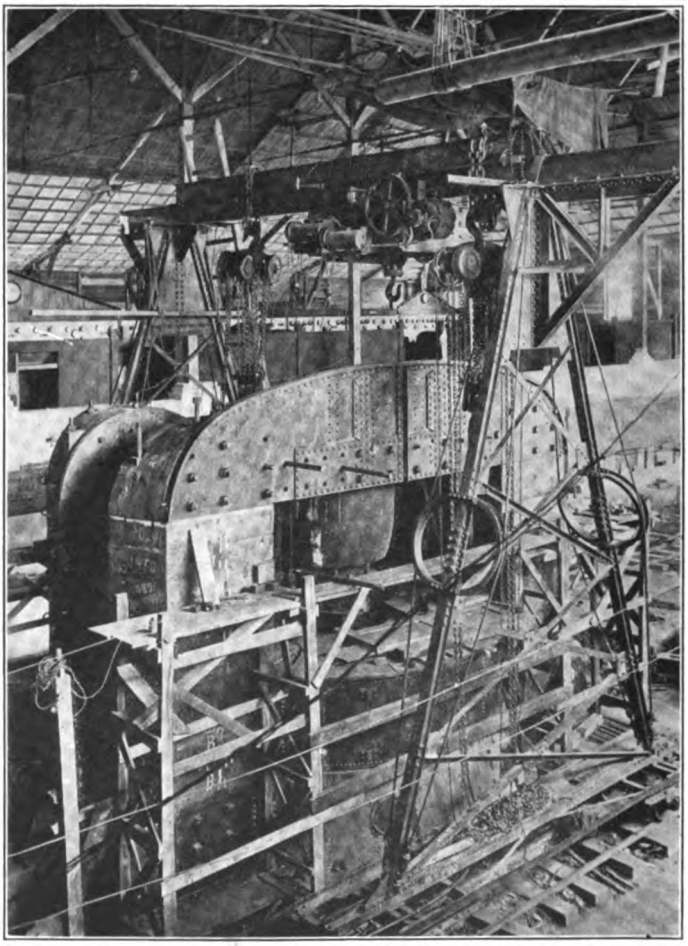

FIGURE 21—Lower Coil in Place and also Upper Magnet Core and Part of Upper Bridge

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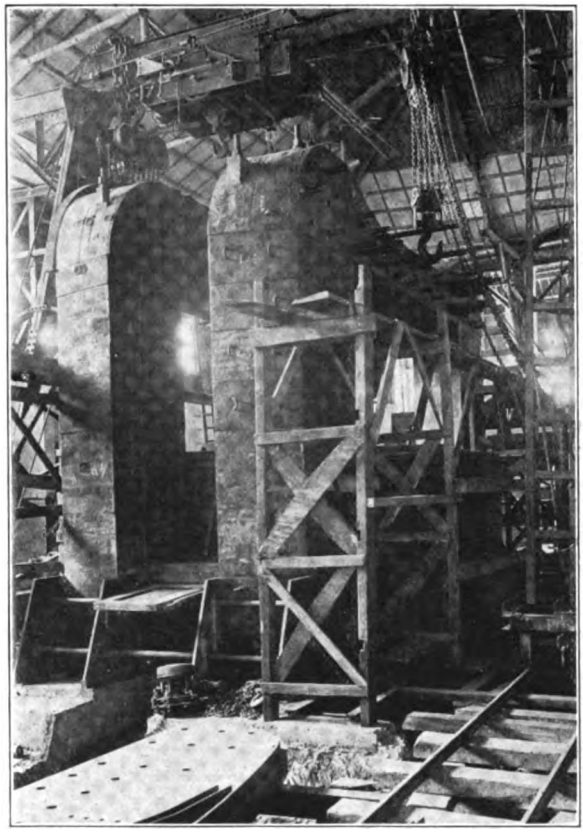

FIGURE 22

of construction was adopted to give the arc mechanical stability, and also to enable the easy insertion of the cathode and anode of the arc. The completed magnetic circuit weighed 260,000 kilograms (260 tons), and appears completed in Figure 23. The erection of the arc was taking place during the photographing of Figure 24, and it will be noted that a graceful tribute has been paid to the radio pioneers, Poulsen and Marconi, by inscribing their names in the tiling on the walls of the station. Arco, Alexanderson Wien, Latour, and Meissner have been similarly honored by inscriptions on the end walls of the station, which are illustrated later in this compilation.

The process of winding the upper magnet coil is shown in Figure 25. This coil is independently excited in order to regulate the field to a value most suitable for the wave length at which operation is being maintained. Figure 26 gives a close-up view

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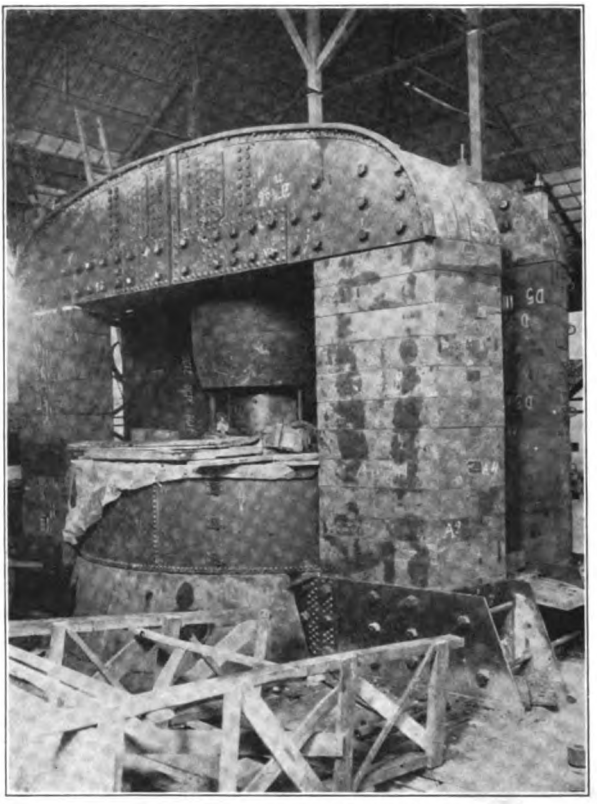

FIGURE 23—Magnetic Circuit

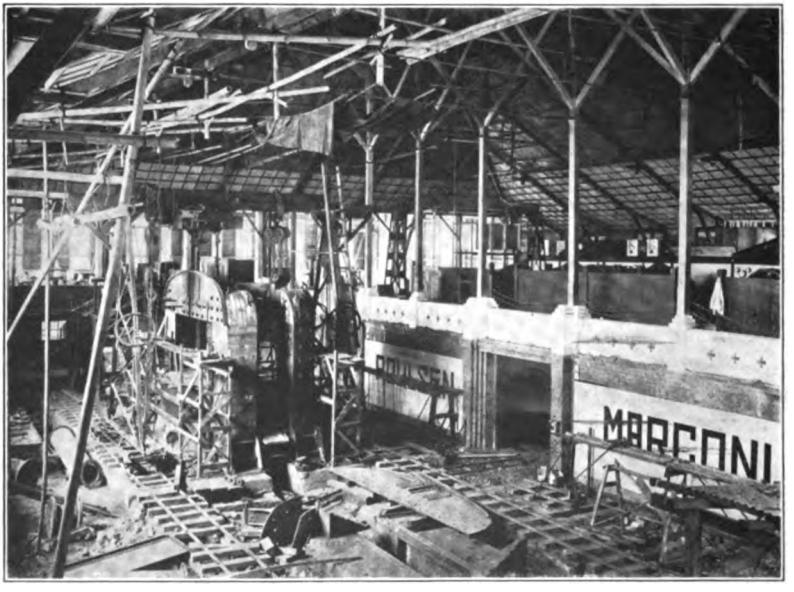

FIGURE 24—Erection of Arc.

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of this upper coil in place together with the interior of the oil tank containing the lower coil. The shape of the conical wrought-iron tip of the lower pole is to be noted. The upper pole tip is of the same shape. By this time, the construction of the arc was

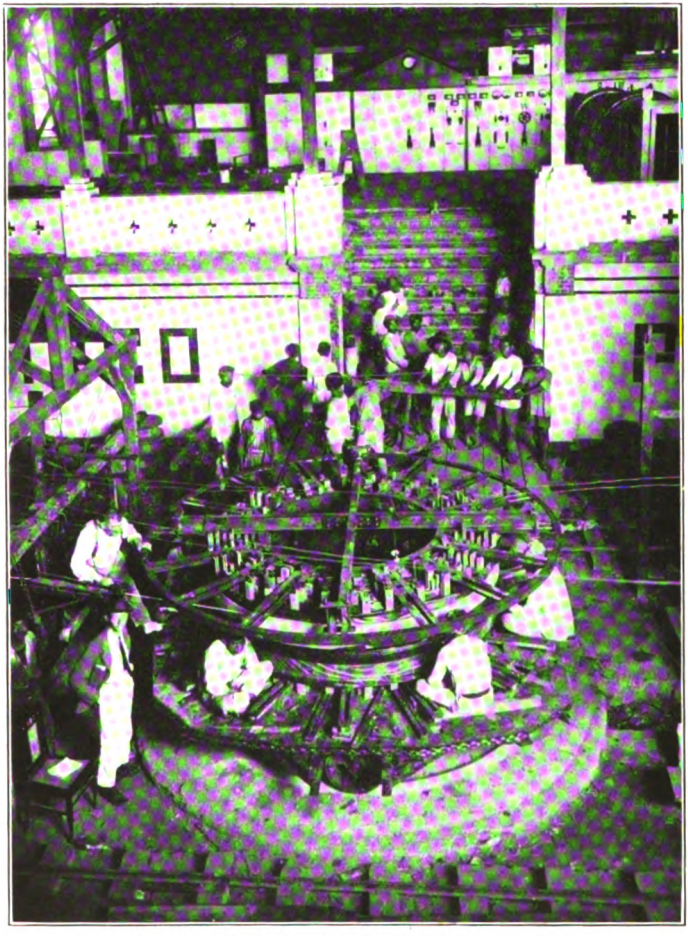

FIGURE 25—Winding of Upper Magnet Coil

proceeding rapidly, and the process of inserting the brass arc chamber (weighing 5,000 kilograms or 11,000 pounds) is illustrated in Figure 27, wherein also appear the magnetic circuits of two auxiliary arcs for military work, each of 200 kilowatts input power. After the arc chamber was in place, the safety valve was fitted into its side as shown in Figure 28. Not long thereafter it was possible to open the station officially, and Figure

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29 shows the memorable occasion on which the first service messages were sent direct by day to the Queen of Holland from the Governor-General of the Dutch East Indies.

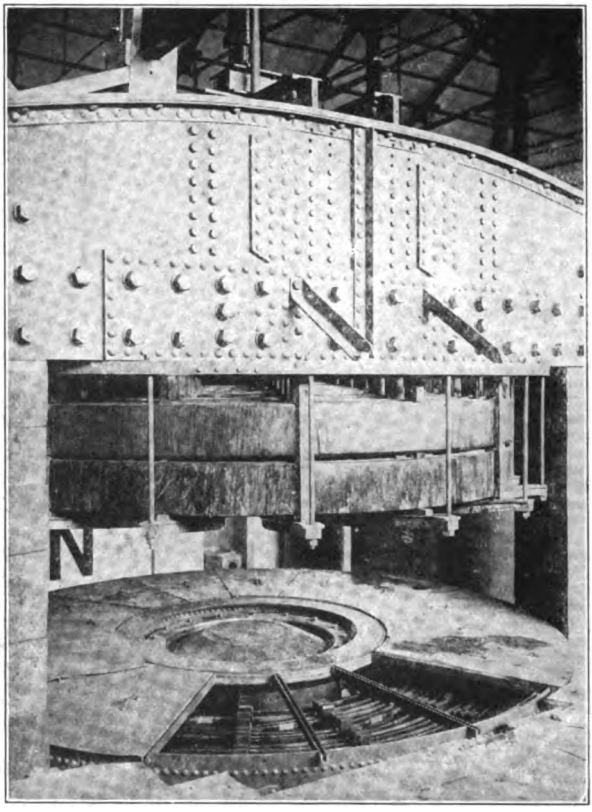

FIGURE 26—Upper Coil in Place, also Interior of Lower Drum

Some of the station auxiliaries merit attention. Cooling water for the circulation system of the arcs and for the fire pumps is run thru filters, and then reaches the arcs. After leaving the arcs, it passes to a cooling pond in the front of the station, whence it is pumped back to the upper reservoir. The cooling at the pond is carried out from horizontal spray pipes. The pipe to the upper reservoir and the tunnel from the reservoir are illustrated in Figure 30. The station is also provided with a plant for producing pure hydrogen for the arcs (to ensure continuous operation without cleaning) and tanks for storing hydrogen and oxygen in sufficient quantities. The cooling reservoir and pumps, while under construction, appear in Figure 31.

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The control of the large radio frequency output of the station presented some problems. A remotely controlled switch had to be provided for selecting whichever one of the arcs was to be connected to the antenna. It is shown in Figure 32. The insulators are of hard rubber, made in the rubber factory at Bandoeng during wartime, and corona helmets are provided over

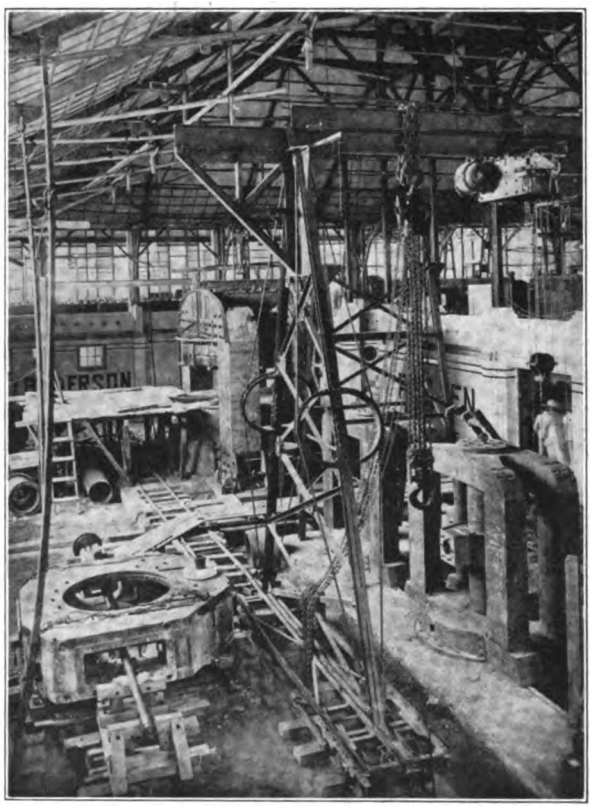

FIGURE 27—Brass Arc Chamber, on Way to Mounting Platform

the switch contacts. The arrangements for the tuning of the station are indicated in Figure 33, which is a photograph of the equipment in process of construction. The loading coil at the right is of copper tubing silver-plated. Two of these coils placed in the main antenna enable covering the wave length range from 7,000 to 20,000 meters. A number of taps connected to a selector switch (remotely controlled) enable tuning to within 1,000 meters by this method. A switch is also provided between the two antenna loading coils to prevent dead-end losses. The

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selector switch in question was under construction at the time this photograph was taken. It consists of a long vertical shaft with a rotating switch arm, with motor drive thru a worm wheel at its lower end. It is mounted on insulators designed to withstand 150,000 volts. For fine tuning of the set, the Rendahl variometer at the left is provided. By its use it is possible to alter the wave length of the station by as much as 1,000 meters

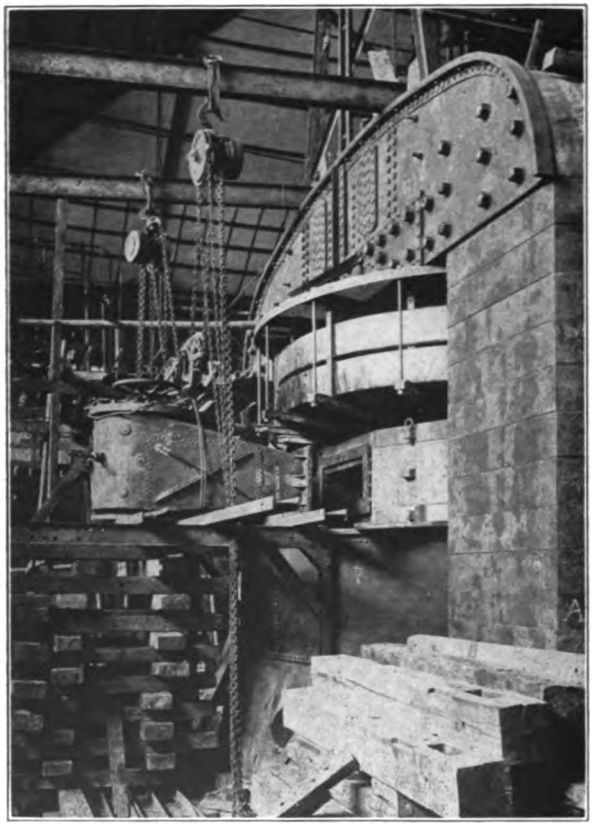

FIGURE 28—Arc Chamber and Safety Valve Being Fitted

while the arc is in operation, which is of importance for a station intended for the special services required in this case. Each arc is provided with its own direct-reading wave meter built by the Lorenz Company. The operator of the station can remotely control the tuning variometer, thereby avoiding even small variations in the wave length of the stations during transmission. In Figure 34 is shown a front view of a portion of the station

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building. The antennas are led out thru a hole in the front of the building, since all forms of porcelain tubes failed in the foggy climate of Java. The down-leads of the antennas are clearly visible in the figures as well as the strings of disc insulators.

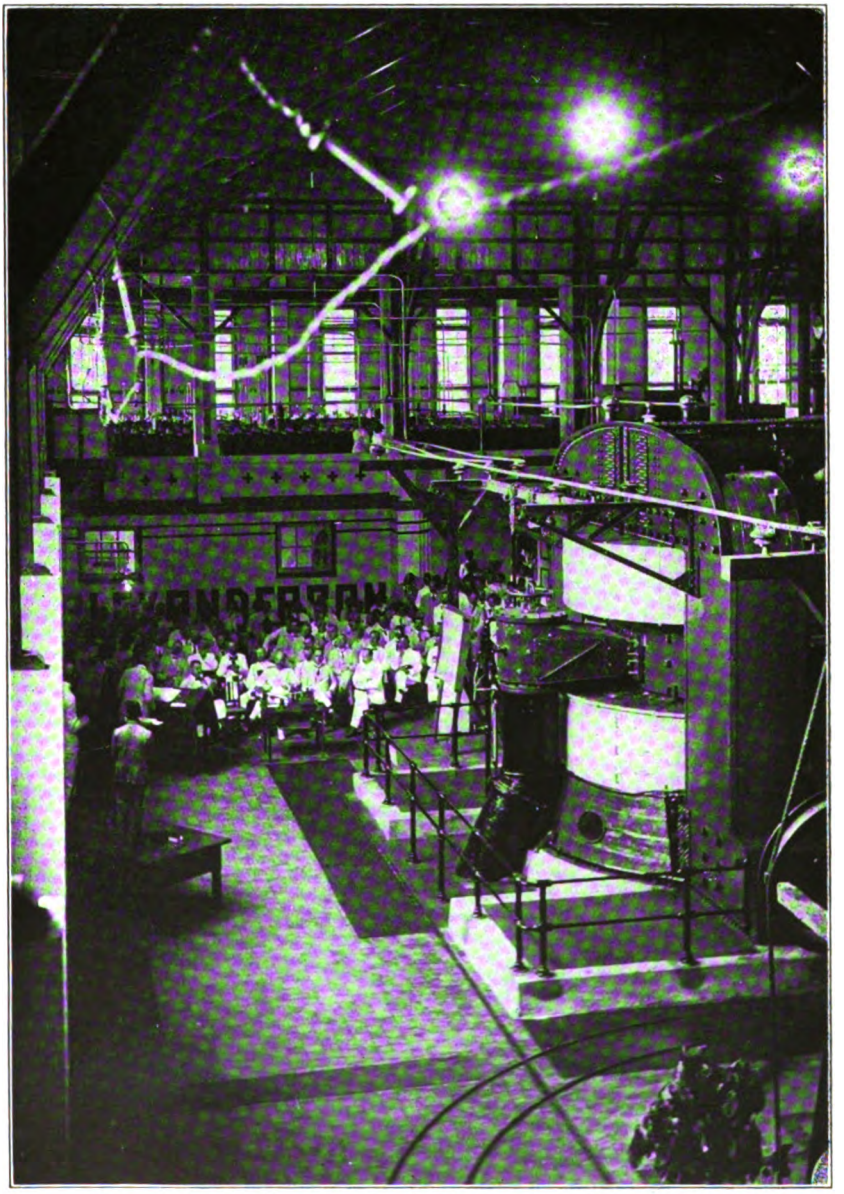

FIGURE 29—Sending First Service Message from Java to Holland

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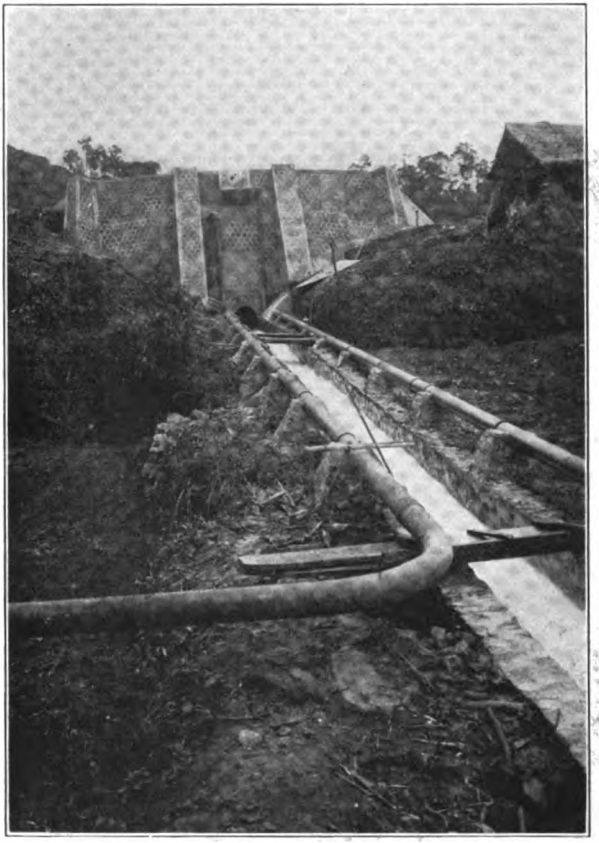

FIGURE 30—Upper Reservoir

FIGURE 31—Cooling Reservoir Under Construction, and Pumps

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These have since been replaced by porcelain insulators. The four towers at the two ends of the building are not purely ornamental. They serve to support the lower ends of the four antennas which it is planned to use for four simultaneous transmissions in the future. The present antenna is supported by five hawsers fixed to the local mountain tops and crossing the ravine. To keep the tension in the antenna constant, the wires thereof are passed over pulleys in the building tower tops and then counterbalanced by weights. At every one of the cross-ravine hawsers, the antenna wires run over large aluminum pulleys on ball bearings, which pulleys are fastened to the lower ends of the corona shields, ending the strings of insulators. The aerial is worked at an effective voltage of 125,000 and shows traces of corona. As it is intended to raise the voltage to 150,000 for full power, a special aerial wire having a ¾-inch hemp core

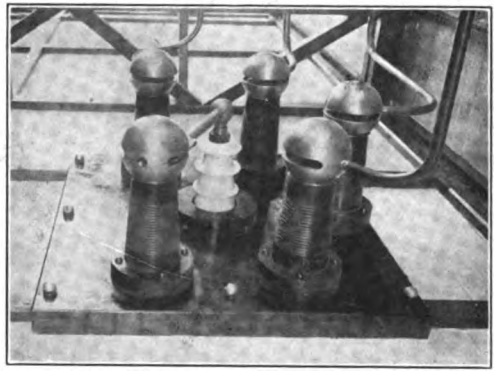

FIGURE 32

will be used in the future. In Figure 34, the Governor-General is shown just leaving the station on its opening day. The cooling pond and the pump appear at the left of the figure.

In addition to the arc installations constructed, erected, and owned by the Radio Department of the Government Telegraph Service, the Telefunken Company was allowed to erect at their own risk in an annex a 400-aerial-kilowatt radio frequency alternator.

The Telefunken Company's installation is shown in Figures 35, 36, and 37. The station is installed on one of the large elevated galleries and is practically a duplicate of those at Nauen, Germany and Kootwijk, Holland. It is still owned by the Telefunken Company. It operates very reliably, though only at manual speed when radiating full power, but is not power-

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ful enough to meet the contractual obligation of providing 24-hour per day commercial service to Holland. It is, in fact, only trustworthy at night, and even then fails absolutely on many nights in the bad season. Because of obligations incurred by the Dutch Government in wartime, it will be purchased. It is an excellent mercantile transmitter for less distant work, such as service to Japan, Saigon, Australia, and Cavite in the Philippines. It serves in this way to collect messages for re-transmission to Europe and can serve during part of the year as a spare transmitter to Europe.

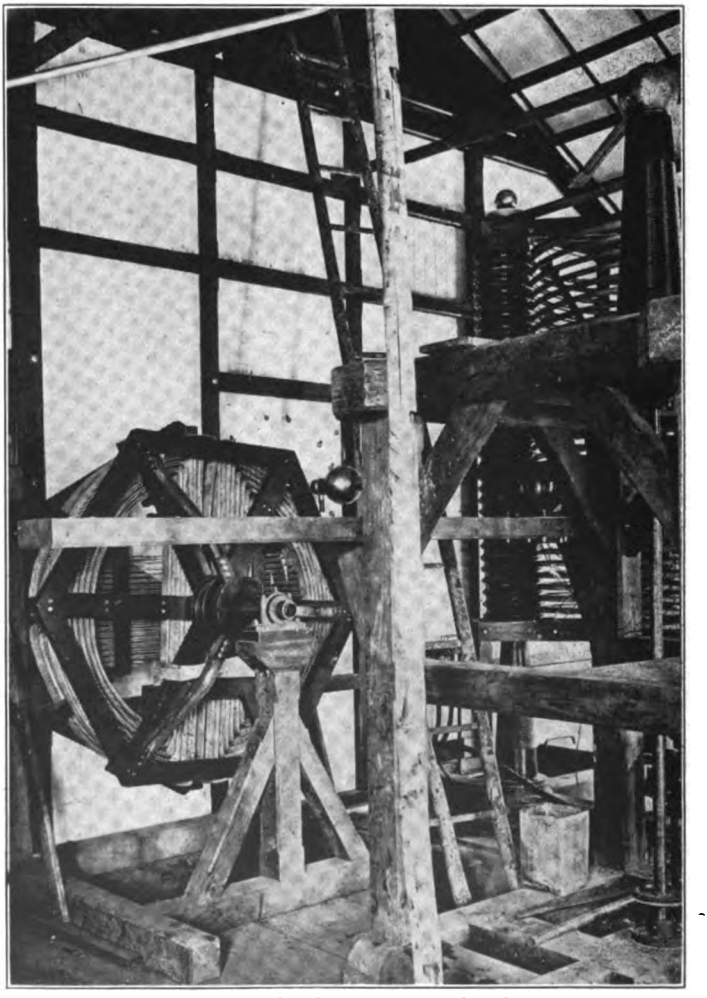

FIGURE 33—High Power Tuning Equipment

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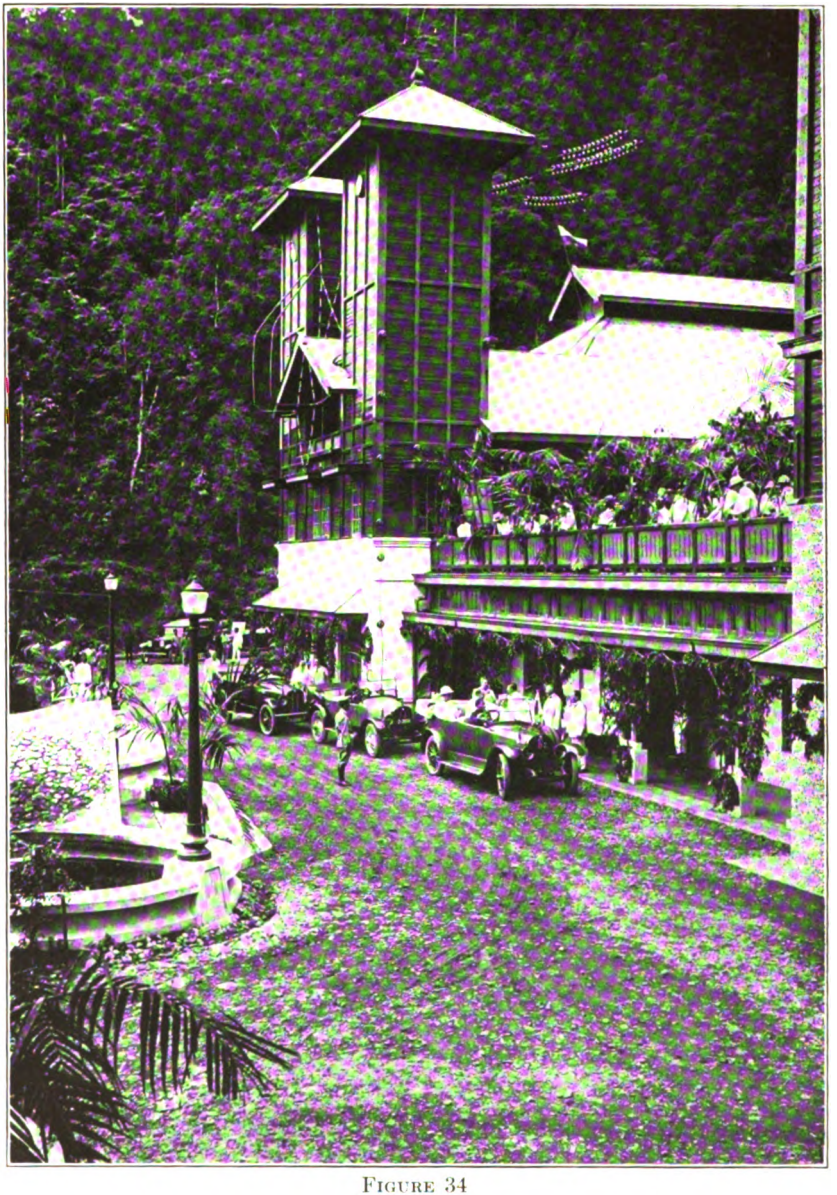

FIGURE 34

In Figure 37, the control relays of the Telefunken alternator set are shown in the front. There is also available a magnetic amplifier for key control of the transmitter which, however, does not function properly at high speeds since the antenna power decreases markedly as the speed increases. In the background of

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this figure are seen the frequency doublers and triplers. The four fixed wave lengths of this set are 7,500, 10,000, 15,000, and 20,000 meters. The first two often provide excellent service to Holland in the night time. The 15,000-meter wave is best in the daytime, but does not give reliable communication on the available power of this transmitter. The 20,000-meter wave can never be copied, either day or night. The results found by experiment are not in accord with the Austin-Cohen transmission formula, since the best wave length for daytime working over a 12,000-kilometer

FIGURE 35

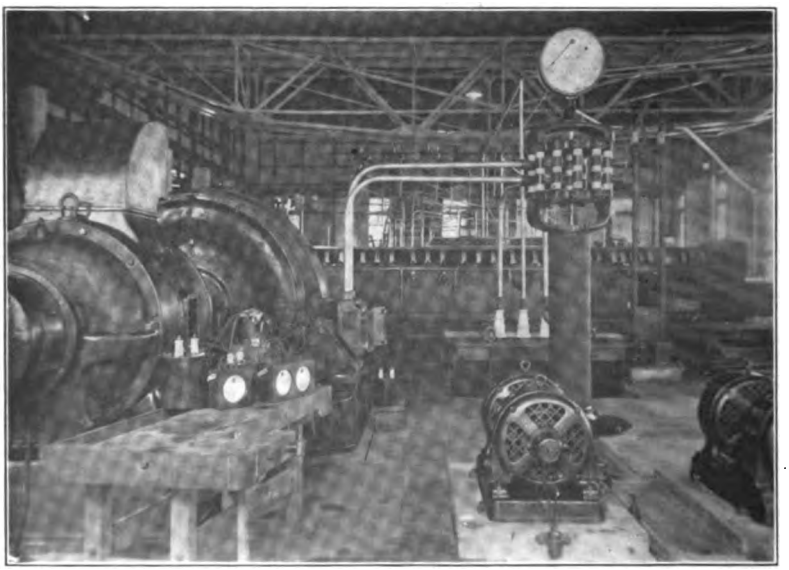

FIGURE 36

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streteh (7,500 miles), from Java to Holland is found to be about 16,000 meters both for the arc and for the alternator, which is several times smaller than the value given by the formula. During complete darkness, the best results are obtained on wave lengths of between 7,500 and 9,000 meters even with a much smaller output. All-year operation requires about 120,000 meter-

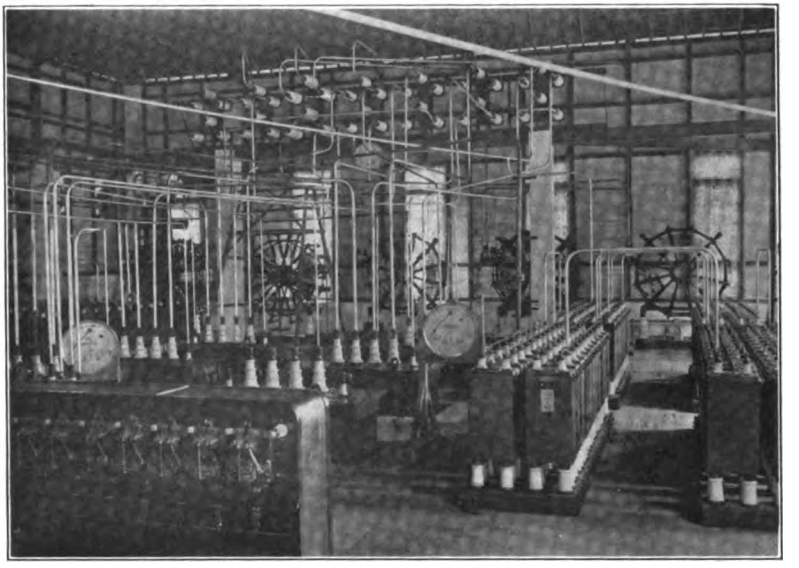

FIGURE 37

amperes on the shorter wave. In full daylight, 250,000 meter-amperes are required on the 15,000-meter wave. During the hours when the transmission path is partly illuminated, the full output of 300,000 meter-amperes is rather below the necessary output for commercial communication. During a recent twenty-four hour test working to Holland, and with the arc furnishing about 200,000 meter-amperes on the 15,600-meter wave at hand speed, communication could be maintained for nineteen hours for plain language text and for eighteen hours with code text. It was possible to deduce from these results that, using the full 300,000-meter-ampere output of the station, twenty to twenty-one hours' daily communication could be dependably secured, which would enable handling more traffic than is now available between Java and Europe. However, it is necessary to await improvements in reception before thoroly satisfactory 24-hour per day communication with Holland can be obtained during the entire year. Of course, the weak 50.000 meter-ampere alternator

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transmitter, working at a long wave length in Holland, gives very poor service in Java thru the strong local static, and is of use chiefly during the night time. This naturally hampers the service from Java to Holland to a considerable extent.

The station being in an isolated location, a repair shop has been provided at the transmitting station. The receiving station is located at Tjangking, and is illustrated in Figure 38. The taller structure in the background is the receiving station, the coil antenna being mounted in the upper story. The lower stories

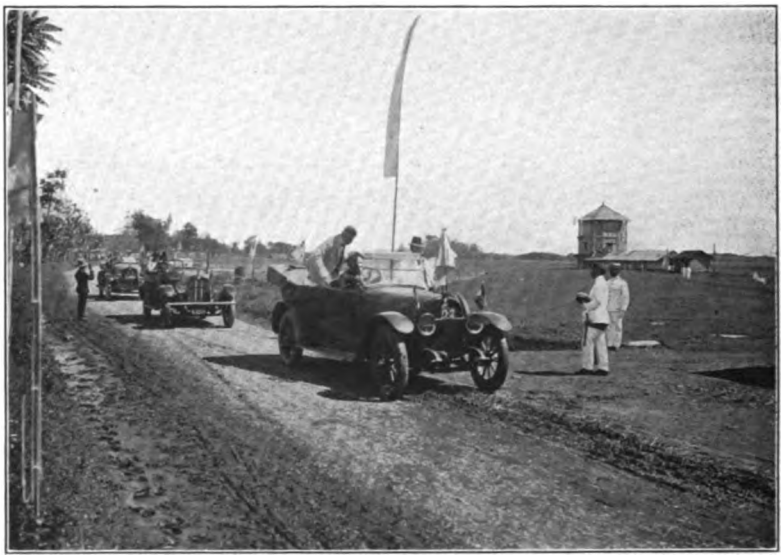

FIGURE 38—Receiving Station, Tjangking

are apparatus rooms completely enclosed by sheet iron. Receivers are provided for four simultaneous services. Either coil antennas, long horizontal wires, or buried wires are used for reception. When the loops are employed, there is no interference whatever from the Malabar station working at a slight wave length separation from the incoming signal, even tho this very powerful transmitter is only 9.5 kilometers (5.9 miles) away, thus demonstrating the purity of the wave of both transmitters. The other structure shown is an exchange or central from which wire lines branch out to the large towns of Java, to the Malabar transmitters, and to the various receivers. This central will be transferred to Bandoeng at a later date. Its present situation has, however, some advantages since no town in Java is a large business center in comparison with large European and American

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cities. Therefore, except for the comfort of the staff, there is no great advantage in having the central located in a large town. Siemens high speed transmitters are used on the circuits leading from the central station to the chief towns of Java. The Malabar transmitters are worked by Wheatstone equipment, even when operated at hand speed, to ensure accuracy and also enable the tape to be inspected before transmission, and to be stored against possible later claims for errors in transmission.

The arc has been operated recently at higher speeds—up to sixty words per minute—reception in Holland being accomplished on a special receiver patented by the Dutch East Indian Government. The transmission is based on an entirely new principle, the keying for compenastion wave being done by a very small pneumatic relay which alters the frequency by only one-half of one part in a thousand in changing from the working to the spacing wave (at hand speed), and from one to one-and-a-half parts per thousand for high speed transmission. Such slight frequency changes are almost unnoticeable in the case of aural reception and are insufficient for copying on all other known forms of recorders, especially at the longer wave lengths. The high degree of frequency constancy obtained at the Malabar arc station is evident from this method of reception. This method of transmission removes the objection to the compensation wave system of keying, enables the use of light and quick-acting control mechanism (which is of great advantage in very high power stations), and even without the use of a code (which limits the operating speed) provides practically absolute secrecy against amateur reception. A portion of a test tape made in Holland on the Malabar reception is shown in Figure 39. This illustrates hand speed and higher speeds up to 45 words per minute.

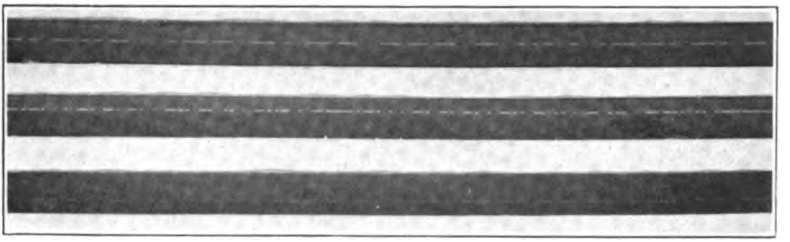

FIGURE 39

The Malabar transmitter PKX operates on the 15,600-meter wave using the arc transmitter or alternator every day except Sunday from 23.40 to 10.40 with brief interruptions when local

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traffic is being handled. Operation is also carried on using the arc on the 13,400-meter wave from 12.40 to 18.40; and from 18.40 to 23.40 the arc or alternator is used mostly on the 7,700-meter wave (altho at times the 13,400- or 15,600-meter wave is used during this schedule). All the above are Greenwich Mean Time. Station PKX can be copied almost continuously at the Radio Corporation of America and Naval stations on the Pacific Coast of the United States. It has been copied during a six-hour stretch at its antipodes—Curaçao—during the previously mentioned 24-hour tests. Communications relative to the reception of this station at remote points, if addressed to the Editor of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, will be promptly forwarded to the administration of the station, who, it is known, will be glad to obtain either graphic records or data on ear reception of this station.

SUMMARY: The construction of the Malabar, Java station of the Dutch East Indian Administration for direct communication with Holland is described. The antenna structure, high power and low power arcs, the alternator transmitter, and the receiving arrangement are considered.

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SHORT-WAVE RADIO BROADCASTING*

By

Frank Conrad

(Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania)

During the year of 1920 the author, for a hobby, maintained a radio transmission schedule with Mr. J. C. Ramsey, of Boston, with the power then available, that is, about 100 watts in the antenna; communication was very uncertain during the summer months and, in fact, had to be abandoned during the particularly unfavorable periods owing to the reduction of the received signal and the increased interference from strays. The radio frequency employed in these transmissions was about 1,200 kilocycles (250 meters). Being aware of the reduction of atmospheric strays on the higher ranges of frequency then in use, and from some experiments in listening to harmonics from other transmitting stations, the author was convinced that there were greater possibilities of improvement in reliability of transmission by increasing the frequency than by decreasing it, as was the general tendency at that time. Accordingly, a series of tests were run between Mr. Ramsey's station (1XA) in Boston and the author's station (8XK) in Pittsburgh. The cooperation of the station of the Massachusetts Institute of Technology (IXM), and that of Mr. R. D. Decker (1RD), of Boston also was enlisted in the tests. These tests were made during the spring of 1921 and consisted of a series of transmissions from each of the Boston stations at various wave lengths and of measurements at Pittsburgh of the audibility of received signals.

The curves in Figure 1 show the ratio of signal strength from the different stations at the various wave lengths operated on. The transmitters were adjusted to give a maximum output at each wave length, with a constant input to the oscillating tubes. Owing to the great variability of the strength of night signals and the small numbers of observarions taken, the results served as only crude approximations, but they indicated some gain of sig-

* Received by the Editor, April 14, 1924. Presented before The Institute of Radio Engineers, New York, April 16, 1924.

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nal strength as the frequency was increased, which, coupled with the great reduction in interference from strays on the higher frequencies, would increase the reliability of transmission with a given antenna power.

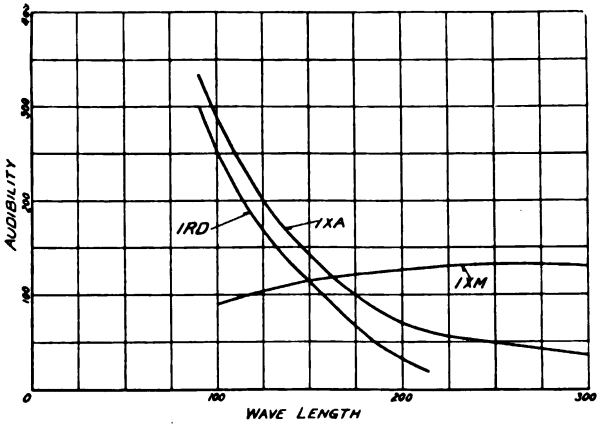

FIGURE 1

To permit of further observations, the equipment at 1XA and at 8XK was remodeled to better adapt it to the higher frequencies. The circuit finally used is seen in Figure 2, which, for the sake of clearness, shows the radio frequency oscillating circuits only. The equipment consists of a coupled circuit in which the local oscillating circuit is coupled to the antenna circuit thru the coupling condenser (C). With this arrangement it is unnecessary to extend the antenna down-lead into the building containing the radio equipment, and as the current in the coupling wire is small, because it represents the energy component only, a comparatively small conductor can be used. This small current minimizes losses due to radio frequency fields. The operation of this transmitting scheme was so satisfactory that it was adopted for the several broadcasting stations operated by the Westinghouse Company. The entire system has been described in a previous paper by Mr. D. G. Little.

As no coupling coils are required, the antenna can be operated at its fundamental without the introduction of loading condensers or inductances. The fundamental of the antenna at 8XK was approximately 2,700 kc. (110 meters), at which frequency it was operated. The fundamental of the much smaller antenna used at 1XA was approximately 5,000 kc. (60 meters),

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and transmissions were carried on from this station on frequencies ranging up to this value. The transmission tests during the following year between these two stations indicated much greater reliability of communication than during previous operation on 1,200 kc. (250 meters). In fact, with the same power which had previously been employed, there were seldom any conditions under which night transmission by straight continuous wave telegraph could not be carried out.

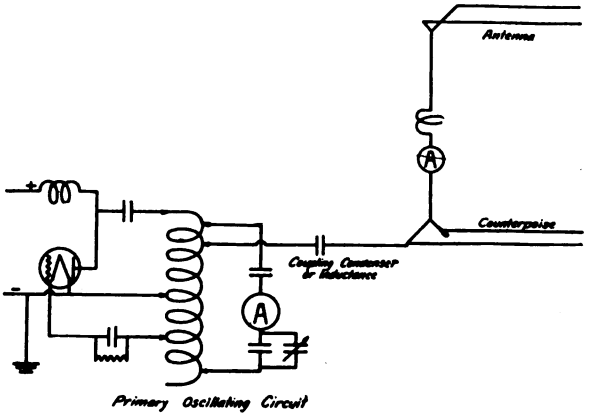

FIGURE 2

As it was evident that the greatest use of these high frequencies would be over the longer distances where communication becomes difficult with the ordinary broadcasting frequencies, it was decided to build a transmitter with sufficient power reliably to cover the range then somewhat uncertainly covered by the broadcasting stations of the usual power; that is, about 500 watts. This transmitter was then built as an adjunct to the transmitting equipment of KDKA, at Pittsburgh, thereby permitting of more continuous observation by transmitting the program of KDKA simultaneously on its regular broadcasting wave and on one of a higher frequency.

The general scheme of the transmitter is the same as that just described. The antenna shown in Figure 3 consisted merely of a vertical pole mounted on the flat roof of one of the factory buildings; the pole being directly connected to the building structure. In order to eliminate high resistance joints in this pole, a number of continuous copper strips were fastened to its surface and at the foot of the pole these strips were carried out over the roof, forming an extended ground. Measurements showed

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that nearly all the current was confined to the copper strips, altho a small percentage was conducted down thru the iron pole to the building frame. The degree of coupling between the antenna and the local oscillating circuit is determined by the point of connection of the leads on the pole or coil and on the impedance of the coupling condenser.

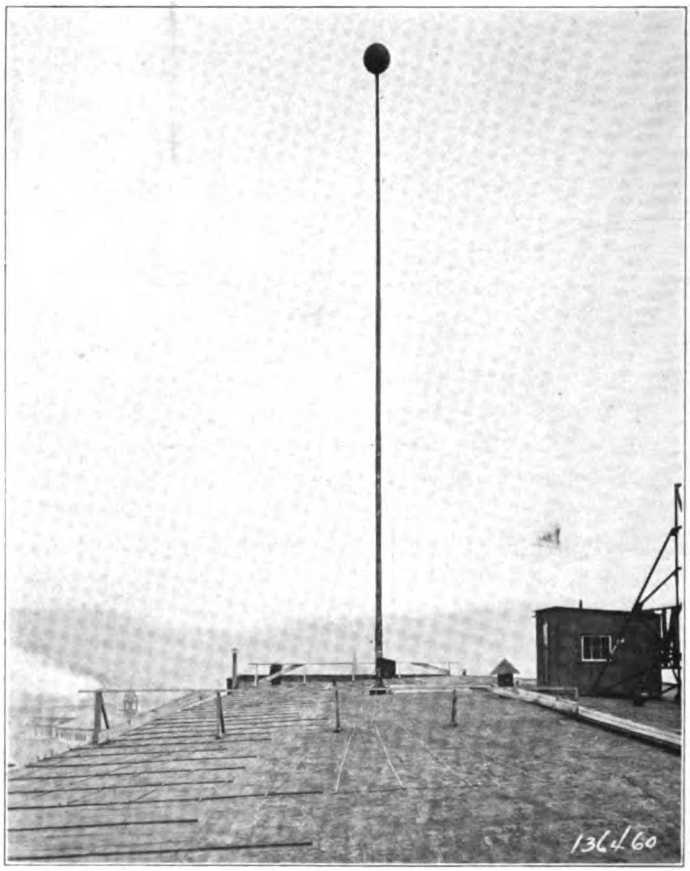

FIGURE 3

In the operation of this transmitter, difficulty was experienced in maintaining a fixed frequency. When operating, the wave would vary thru a range of frequencies of as much as 1,000 cycles, so that it was very difficult to carry on telegraphic reception by the heterodyne method or to receive telephone signals on a sharply tuned set. This variation of frequency was appar-

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ently due in part to changes taking place in the various circuits of the building, which acted as an extended ground. Any alteration of circuits in the interior of the building was found to have a slight effect on the resonant frequency of the antenna.

To overcome this difficulty, the vertical-pole antenna was replaced by the conventional form of inverted "L" with insulated counterpoise, the counperpoise being placed about 12 feet (3.7 m.) above the building. This arrangement was found to overcome the abrupt frequency shifts, but there remained the general irregular fluctuations of this frequency due to various causes, such as the vibration of connecting wires or the variation of plate supply voltage. To overcome the effects of vibration, the inductances were wound from strap on heavy frames, and all connecting conductors were made from copper tubing rigidly supported on insulators. The small remaining vibration due to machinery in the building was eliminated by suspending the whole transmitting structure from a set of springs. To improve the regularity of the supply voltage, power was supplied to the transmitter thru separate lines run direct to a special generator in the powerhouse. The effect of these precautions was greatly to improve the constancy of frequency.

For telegraphic communication it was found preferable to signal by key modulation or the so-called spacing wave rather then by totally interrupting the antenna power; the use of the spacing wave eliminating key clicks in receiving sets tuned to broadcasting waves. The spacing wave was about 500 cycles removed from the signal wave, this shift being brought about by shunting a small capacity across one turn of the primary tuning inductance. The amount of energy in this additional shunt circuit was so small that it could be controlled by a relay similar to the ordinary telegraph relay. In the final form of relay, no contacts were used, the change in capacity shunted across one of the turns being brought about by the movement of a metal plate attached to the relay arm, and it was found that a plate about two inches (5.08 cm.) in diameter moving thru a distance of one-eighth inch (3 mm.) was sufficient to control a transmitter delivering 1 kilowatt to the antenna.

This transmitter was equipped with four 250-watt air-cooled oscillators and five modulators of similar type, the modulator tubes being disconnected when the set was operated for telegraphy. The four oscillators delivered approximately 800 watts to the antenna circuit.

The first transmission experiments were carried on between

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Pittsburgh and Cleveland, attempts being made at Cleveland to pick up the high-frequency signals and relay them thru a small broadcasting set installed there. On the night of October 27, 1922, KDKA's program was first relayed in Cleveland from 8:00 to 10:00 P.M. Cleveland is only about 110 miles (176 km.) from Pittsburgh by air line, but lies in one of those peculiar radio areas where reception of signals from Pittsburgh is comparatively difficult. It was found that the signals received at Cleveland from this high-frequency transmitter were very much louder than those received from the regular broadcasting transmitter, having about the same power output. It also was found that the signals were nearly as loud during the daylight hours as at night. However, the following peculiarity was noted: When operating at 3,300 kilocycles (91 meters), the signals were somewhat louder at night, but when operating at 3,750 kilocycles (80 meters), the signals were louder during the daylight hours than during the night. Interworks communication was carried on for a short time with the messages being transmitted by telephone on a wave length of 80 meters during the daylight hours, but changed to 90 meters when darkness came on.

At first serious difficulty was experienced in obtaining good quality from telephonic transmission, the signals received being very much distorted. This distortion was found to be caused in part by frequency changes caused by variations of average plate current by modulation, and to a peculiar effect of high-speed fading, which at times became so marked as to give an audible noise in the receiving station even tho a perfectly continuous unmodulated wave was being radiated from the transmitter. The distortion due to changes of oscillating constants in the transmitter caused by modulation was reduced by careful adjustment of the set. However, the distortion due to causes existing in the intervening space could not be entirely overcome, but it was minimized by the employment of a comparatively strong signal; a hand adjustment was employed to compensate somewhat for the changes due to slow fading.

During 1922 some development work was made on a special type of transmitter for operation on 3,000 kilocycles by F. W. Dunmore of the Bureau of Standards. Transmission experiments with this equipment at Washington and the set at 8XK indicated a daylight signal strength which was only slightly less than that of the night signal.

Reports received from various sections of the country on comparative signal strength of the two waves from KDKA, indi-

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cated the possibilities of establishing a broadcasting system which would cover the entire country with a comparatively high degree of reliability. To carry out this scheme, it was thought advisable to install a transmitter about midway between the Pacific Coast and Pittsburgh to act as a repeater station. Hastings, Nebraska, was selected as the location for this station, its distance from Pittsburgh not exceeding that which our tests indicated could be covered with fair reliability.

Analysis indicated that an antenna input of possibly 10 kilowatts would be required for reasonable reliability. Accordingly, two complete new transmitters were built; one being installed at East Pittsburgh and one at Hastings. The tubes used in these sets are water-cooled 10-kilowatt capacity, but to allow an ample margin two tubes are operated in parallel as oscillators and two as modulators. The plate voltage is eight thousand volts. As might be expected, this increase in power brought with it a train of miscellaneous difficulties one of them being the construction of a condenser which could handle the large kilovolt-amperage required. Mica condensers of conventional design which had been used on the previous sets were found to be very unsatisfactory for the higher power, owing to their greatly reduced kilovolt-amperage capacity at these higher frequencies. A type of oil condenser was developed which, altho simple and comparatively inexpensive, gave very satisfactory results. This condenser is shown assembled in Figure 5 and in partially disassembled condition in Figure 6.

The type of inductance used which is shown in Figure 7 consists of a helix of thin copper ribbon wound on Pyrex glass blocks supported on a rigid frame of vacuum-impregnated wood. The assembled transmitter units comprising rectifiers, modulator, and oscillator are shown in Figure 8. The scheme of connection is shown in Figure 9 and the arrangement of the antenna in Figure 10.

The transmitter at East Pittsburgh was designed to operate only within the frequency range of 3,000 to 3,600 kc., but the set at Hastings was designed to operate at a frequency within this range or at a lower one, within the regular broadcasting range. This arrangement permitted operation under a scheme in which the short wave signals being transmitted from KDKA could be received in Hastings, and from there re-transmitted on the regular broadcasting wave (800 kc.). With the power available the signals from this station would, during the favorable season of the year, be received over the entire United States with a good signal —stray ratio.

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It was possible also to re-transmit from Hastings on a high frequency wave which was necessarily spaced sufficiently from the incoming wave of station KDKA to prevent feed-back troubles in the receiving equipment used at Hastings. Advantage then could be taken of the better transmission efficiency of the high frequency wave when it was desired to reach the Pacific Coast with a signal capable of again being relayed.

FIGURE 5

The receiving atation used at Hastings developed some troubles which, altho not entirely unforseen, yet were of considerably greater magnitude than had been expected.

Hastings is a small city, the population being less than fifteen thousand, and contained no feature which would ordinarily be expected to set-up interfering strays. When the high frequency receiving equipment was installed in the central part of town, it was found that a fairly complete log could be obtained of the operation of nearly every X-ray, violet ray, and massage machine, as well as the number of street lights and elevator motors that were being operated. This, of course, does not imply that satisfactory audible signals could not be received, but it did indicate that for repeating purposes a considerably higher signal-stray ratio was required.

The signal distortion which was noted in preliminary experi-

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ments at Cleveland was experienced, as might be expected, to a more marked degree at Hastings, but was subject to great irregularity of intensity; some nights the received signals being practically undistorted while on other nights speech would be unintelligible. The greatest factor in bringing about this distortion was apparently the peculiar high-speed fading and fluttering of signals. From observations of fading, phenomena as related to ordinary broadcast waves, it is known that the phase of the fading cycle is not constant over the whole area within which a signal can be received; but, in fact, may show a considerable change at receiving stations located within a few wave lengths of each other.

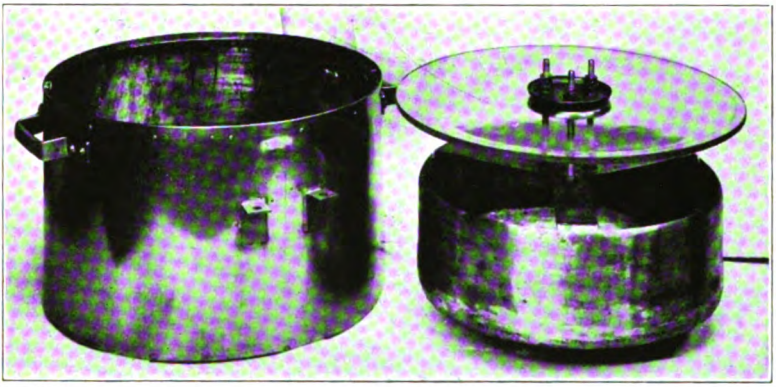

FIGURE 6

To take advantage of a possible overlapping of the fading cycle, two receiving stations were installed, the distance between these stations being about three miles (4.8 km.). As they were not on a line intersecting the transmitting station, the actual difference in distance to Pittsburgh was about two miles (3.2 km.). This arrangement incidentally solved the local stray problem as the receiving stations were installed in the outlying farm districts beyond the disturbing radius of the town's activities. The receiving equipment, a diagram of which is shown in Figure 11, consists of a conventional coupled circuit tuner, detector and two stages of audio amplification. The inductance and capacity elements, which are connected in parallel with the primary tuning element, constituted wave traps which served to eliminate interference from the local transmitter. To provide ample output and to minimize distortion, two tubes are used in the second audio stage. These are connected in push-pull fashion. The out-

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put transformer of the amplifier is wound to the proper step-down ratio to suit it to the connecting lines between the receiving and transmitting station. At the transmitting station the incoming lines from the two receiving stations may be paralleled and connected to the input terminals of an additional amplifier which steps up the received voltage to the value necessary for control of the modulator tubes.

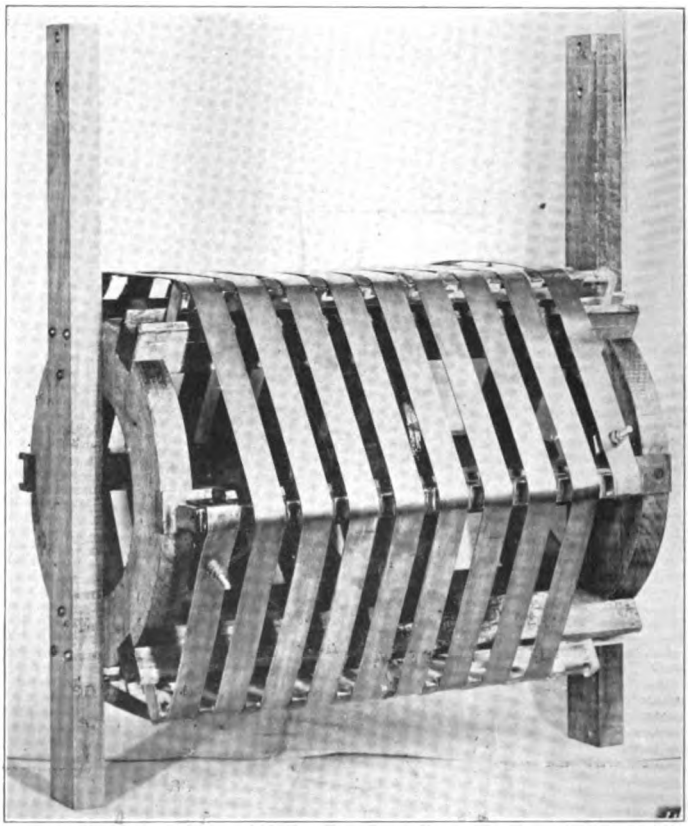

FIGURE 7

To provide for remote control of the receivers, a connection is made to the center of the line transformer windings whereby direct current may be sent over two lines in parallel in connection with a ground return. This serves to operate a relay at the receiving station for controlling the receiving tube filament circuit; the receiver having been previously tuned to the incoming signal wave.

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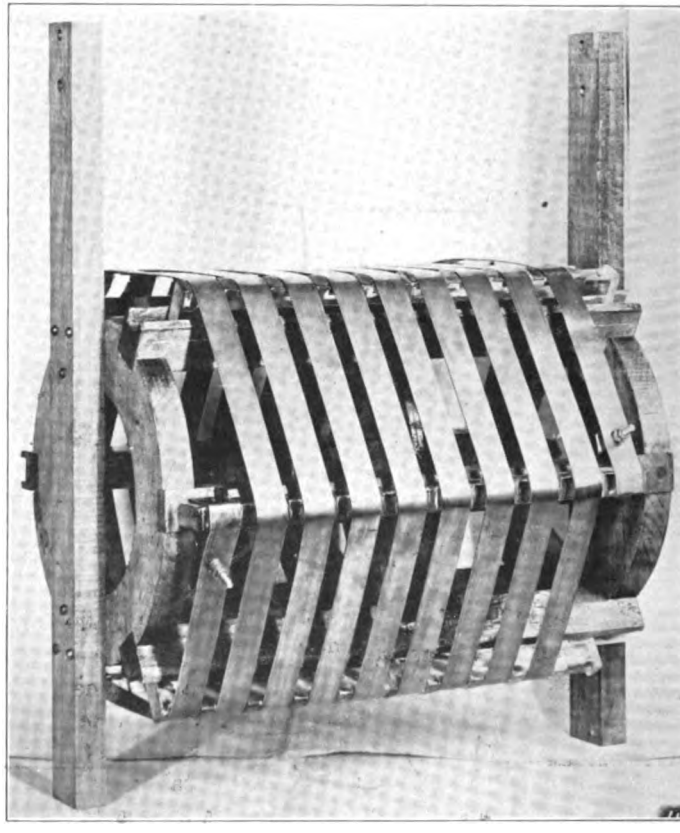

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The use of multiple receiving stations at certain times were found to reduce materially the signal distortion. At other times, however, it seemed to have no effect, so that the arrangement, altho effecting an improvement, could not be considered a solution. It is hoped, however, that investigation of the best separating distances between the receiving stations may produce further improvement.

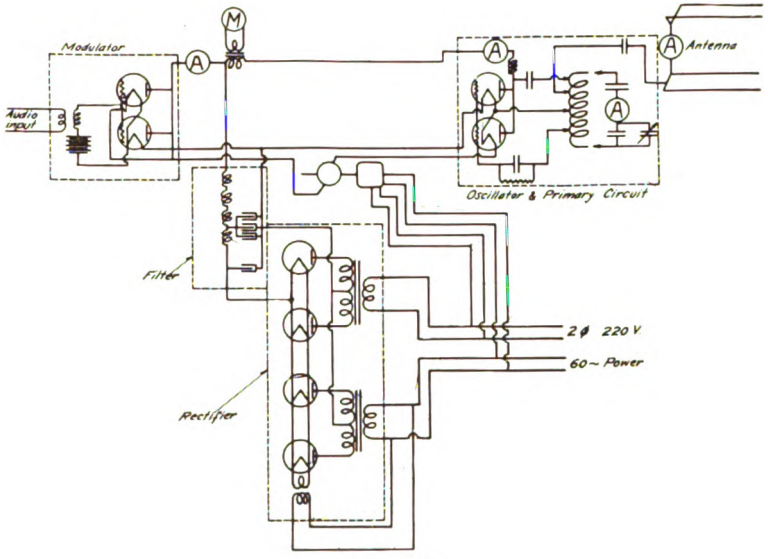

FIGURE 9

From experiments, which have been carried out, the most promising possibilities of undistorted transmission appeared to be in the selection of the particular frequency which will give the minimum fading distortion; this frequency value possibly varying from day to day.

The general trend will be toward a higher frequency, the limitation in this direction being those incident to the building of transmitting equipment of the capacity required.

Experiments which have been carried on heretofore on what might be termed ultra-high frequencies, were at comparatively low powers. The low power units, however, except in certain particulars such as directivity, would not compete in transmitting ranges with even the smaller conventional wave broadcast equipments.

The upper frequency limit of the sets now installed at Pittsburgh and Hastings is about 3,800 kc. (79 meters), the principal

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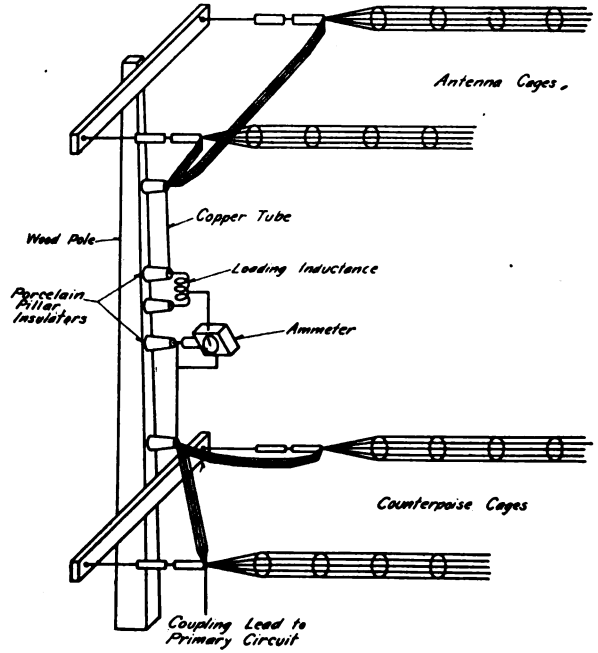

FIGURE 10

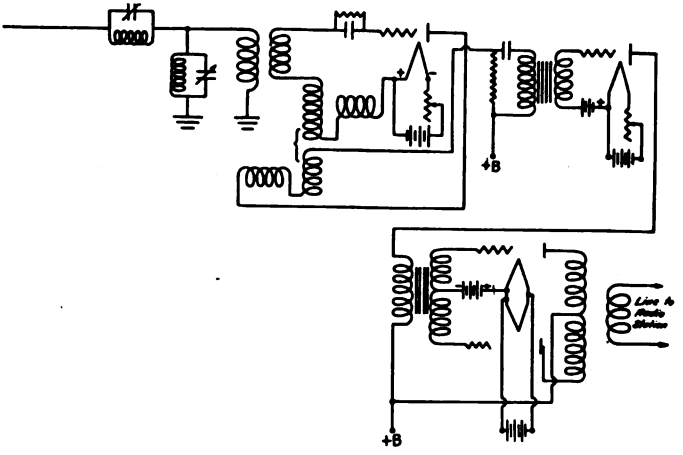

FIGURE 11

limitation being the current-carrying capacity of the grid lead-in connections and the heating of glass parts due to radio frequency electrostatic fields. This limitation will be appreciated when it is known that the grid charging current at about thirty-eight hundred kc. is approximately ten amperes.

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New tubes are now in the course of construction which should permit an increase of frequency to possibly 6,000 kc., at which value when under full power, the set, like the traditional "one-horse shay" probably will reach a general break-down limit.

So far, no accurate measurements of absolute signal strength have been made, but there has been taken a rough survey of the country, to determine the signal strength at various locations. A number of small receiving sets consisting of a tuning element detector and one stage of audio amplification were sent to various parts of the United States, England, and Canada. These sets were used with the same general type of antenna. The antenna used had a total length of 15 ft. (4.6 m.) and was usually strung indoors. With this receiving equipment, telephone transmission was intelligible in nearly every location with audibilities ten times and upwards.

The reduction of signal strength with the increase of distance from the transmitter differs widely from that which might be expected. Within a few miles of the transmitter the signal strength from the high frequency transmitter is about one per-cent of that from the regular broadcasting wave transmitter. There are areas a few hundred miles distant where this approximate ratio is reversed.

As we approach the limits of range of the regular broadcasting transmitter, we are all familiar with the phenomena of good reception on some nights and no detectable signals on other nights. At these distances the signals from the high frequency set are markedly uniform, reception in general being possible on the receiving set just described. Reports received from the sets installed in England indicate that the intelligibility of received signals is determined in a great measure by the degree of distortion due to fading, rather than by actual signal strength.

It seems almost a flight of fancy to state that signals from East Pittsburgh are picked up in England on the two-tube dry cell detector used with an antenna from 10 to 15 feet (3 to 4.6 m.) long and the whole inside of a building, but it is a fact.

The areas of intense signal strength brought out unlooked-for defects in the selectivity of receivers adapted to receive the usual broadcast wave. At first, when we started operating the high frequency transmitter at KDKA, comments came in on the broadness of the KDKA wave and each step of power increase of the high frequency transmitter would open up new areas where the general comments would be "The KDKA wave was all over the scale." This phenomena applied to all types of tuners. The

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apparent lack of selectivity of the tuner was due to the fact that sufficient signal voltage was being induced in the antenna to give a response of the detector independent of any building-up due to resonance. This condition was illustrated by several experimental transmissions sent out from Hastings. This station transmits two local musical concerts each week on the regular broadcasting wave. The transmitting frequency of several of these concerts was shifted to the high-frequency wave; that is, in place of being transmitted on the regular wave of 880 kilocycles, they were transmitted on a wave of approximately 3,000 kilocycles. A great number of letters were received from listeners, indicating a quite general reception of these high-frequency transmissions, altho the receivers were in most cases the usual type of broadcasting receivers, having a frequency limit of about 1,500 kilocycles. However, the almost universal comment was that the reception was particularly good as regards signal strength but that the station could not be tuned out, it being equally loud on all parts of the tuning scale. Some of the letters received from as far as 1,000 miles (1,600 km.) away indicate reception on receivers of the simple crystal detector type.

The simplest method of eliminating interference from these high frequencies is the use of a wave trap, or resonant circuit, tuned to the high-frequency wave and connected from antenna to ground terminals of the receiver. This trap, if made with a comparatively low value capacity unit, will have an inappreciable effect on signals of ordinary frequencies.

The results obtained so far indicate that, altho there is considerable further development required, the use of these higher frequencies will be a decided forward step in extending the range of broadcasting stations. For comparatively short distances there seems no particular advantages to be gained over the normal waves other than the possibility of increasing the number of communication channels. However, it is in the possibility of greatly extending the broadcasting radius that the greatest promise lies, the first application being the broadcasting of items of international interest and importance. Several transmission schedules have been carried out thru the East Pittsburgh-Hastings transmitters which indicate these possibilities.

On November 22, 1923, a talk given at Pittsburgh by Mr. E. H. Sniffin was received at a meeting of the National Electric Light Association being held at Salt Lake City, the transmission circuit being from KDKA to KFKX (Hastings, Nebraska), on 3,000 kilocycles; from there re-transmitted at 1,050 kilocycles,

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at which frequency it was received at Salt Lake City. This event possibly marks the first regularly scheduled long distance relay transmission.

Tests were also being carried on between an experimental receiving station located at the Metropolitan-Vickers Electrical Company's Works in Manchester, England, on the high-frequency waves transmitted from KDKA. So successful were preliminary tests that it was decided to institute a special service for English listeners. Arrangements for the event occurred near the close of the year 1923, so it was decided to hold the actual broadcasting of the first English program from KDKA until the New Year. Accordingly, New Year's Eve, Mr. H. P. Davis, Vice-President of the Westinghouse Company, broadcast a New Year's greeting from the "Pittsburgh Post" Studio. This greeting was transmitted at 7:00 P. M., Eastern Standard Time, which, because of the difference in time, was 12 o'clock in Great Britain. This message was received on the short-wave receiver, was transmitted from station 2ZY of the Metropolitan-Vickers Company and was simultaneously transmitted from the other seven stations of the British Broadcasting Company. This was the first pre-arranged regular broadcasting sent to England for re-broadcasting, from KDKA.

All of you, probably, are familiar with the transmission of the dinner held by the Massachusetts Institute of Technology Alumni at New York City on March 7. By several relay stages the speeches and music at this dinner were received over the greater part of the English-speaking world. At this time the signals were transmitted from the Waldorf-Astoria Hotel, where the dinner was held, to the Radio Corporation's station (WJZ) at New York and the General Electric Company's station (WGY) at Schenectady by direct wire connection. At Schenectady the signals were transmitted by radio on their regular broadcasting wave, as well as on the high-frequency wave of approximately 2,900 kilocycles. Both these waves were received at Pittsburgh and re-transmitted by KDKA on 920 kilocycles and on 3,000 kilocycles. At Hastings the 3,000-kilocycle wave was received and re-transmitted at 1,050 kilocycles. At Oakland, California, both the 1,050 kilocycle wave from Hastings and the 3,000-kilocycle wave from East Pittsburgh could be received. This latter wave was received with the greatest signal intensity and was therefore used in re-transmitting on their regular broadcasting wave of 900 kilocycles. In addition to this, the 3,000-kilocycle wave from East Pittsburgh was received in England

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and re-transmitted thru several of the British Broadcasting Company's stations. The success of this undertaking serves to show the possibilities of this high-frequency relaying system.

SUMMARY: A brief history of short wave broadcasting as carried on under the direction of the author is given, together with the difficulties encountered and the results obtained. The transmitter and receiver employed experimentally for repeating KDKA's programs from KDPM at Cleveland are described. The high power short wave transmitter at 8 XS (Pittsburgh) is described particularly in regard to the radio frequency condensers and inductances required. A description of the receiver and transmitter at the Hastings, Nebraska, repeating station KFKX is given.

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DISCUSSION

Greenleaf W. Pickard (by letter):* Mr. Conrad's timely paper is one of deep significance to radio engineers. For a quarter century the higher transmission frequencies have remained in innocuous desuetude, while the infra-red of the radio spectrum has been investigated and utilized. But now that the ultra-violet had been harnessed, we may expect a rapid extension of our knowledge in this direction, and the certain addition of many useful frequency channels.

From the time that KDKA at Pittsburgh began its high frequency broadcasting I have been an interested observer, and for some time past a recorder as well. When WGY at Schenectady began a similar transmission, I was able to make reception comparisons at my laboratory at Newton Centre, Massachusetts, from transmitters at nearly equal frequencies at distances of 760 and 225 kilometers (475 and 145 miles), respectively. And altho a fair picture of broadcast transmission at 3,000 kilocycles (100 m.) would require a far more extensive investigation than mine, I feel that certain of my results may be of interest.

Thru the courtesy of the Westinghouse Company, KDKA radiated for me at its two frequencies, 920 and 3,000 kilocycles (326 and 94 m.), thru dawn and sunrise of January 15th and 16th, 1924, and in Figure 1 is shown the record of these two transmissions.

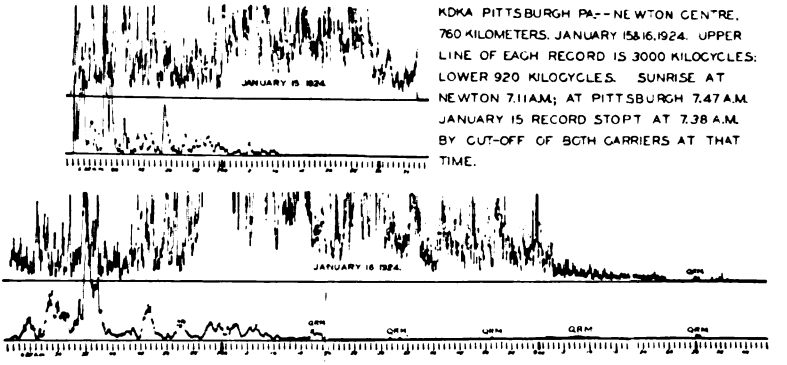

FIGURE 1

On these two mornings sunrise at Newton was at approximately 7.11 A. M., and the records show that this was an epoch time for the 920-kilocycle field, which then sank to its normal

*Received by the Editor, May 6, 1924.

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low daytime intensity. But the 3,000-kilocycle field continued without material change thru sunrise at both Newton and, thirty-six minutes later, sunrise at Pittsburgh. At 8.03 A. M. the high frequency field fell off markedly, and finally at 8.24 A. M. it reached its normal low daytime intensity, an hour and thirteen minutes after the 920-kilocycle field.

Thru the day both the 920- and 3,000-kilocycle fields are too near the disturbance level for satisfactory recording. But two or three hours before sunset at Newton the 3,000-kilocycle field begins to rise, and good reception can usually be obtained at this frequency for nearly two hours before the 920-kilocycle field has risen sufficiently above the noise background for reception. So far as Newton Centre is concerned, KDKA's high frequency radiation has therefore added at least three hours of possible reception to the day.

WGY at Schenectady, now radiating at 790 and 2,800 kilocycles (380 and 107 m.), and at less than one-third the distance of KDKA from Newton, is well above the noise background at any hour of the day or night at its higher frequency. But the 2,800-kilocycle field shows severe and rapid fading, both by night and by day. Figure 2 fairly represents the normal daytime fluctuation of this station at its higher frequency. The normal 790-kilocycle transmission from this station, as I have shown in a recent paper,[1] shows practically no daytime fading in Newton.

WGY MARCH 20&16,1924. MORNING AND AFTERNOON AT 2800 KILOCYCLES.

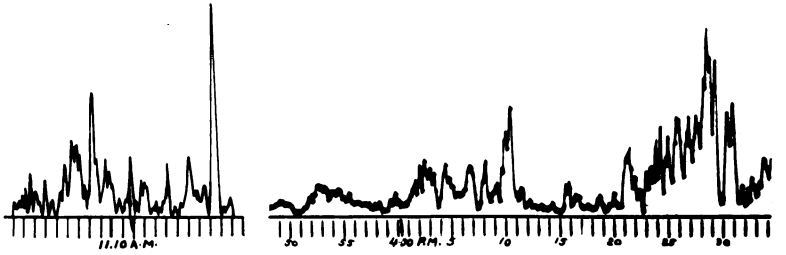

FIGURE 2

Another interesting effect of high frequency transmission is its penetration of steel frame buildings. The other evening, in

[1] "Short Period Variations in Radio Reception," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 12, number 2, 1924, Figure 7.

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my room on the fifteenth floor of the Commodore Hotel in New York, I was able to receive KDKA at 3,000 kilocycles with excellent intensity on a small indoor antenna, whereas, with approximately the same amplification, there was no trace of the 920-kilocycle wave; an indoor field difference of certainly a hundredfold. A similar difference was also observed with WGY.

Night-time reception of KDKA at Newton at 3,000 kilocycles is frequently marred by a severe and curious impairment of quality, usually accompanied by a rushing or roaring sound. Similarly, in receiving WGY either by night or day at 2,800 kilocycles, there is an even more pronounced and frequent distortion. By using a string galvanometer for recording, I have discovered that the distortion is caused by an ultra-rapid fading, the fluctuation frequently being at audio frequencies. In Figure 3 is shown one of these records.

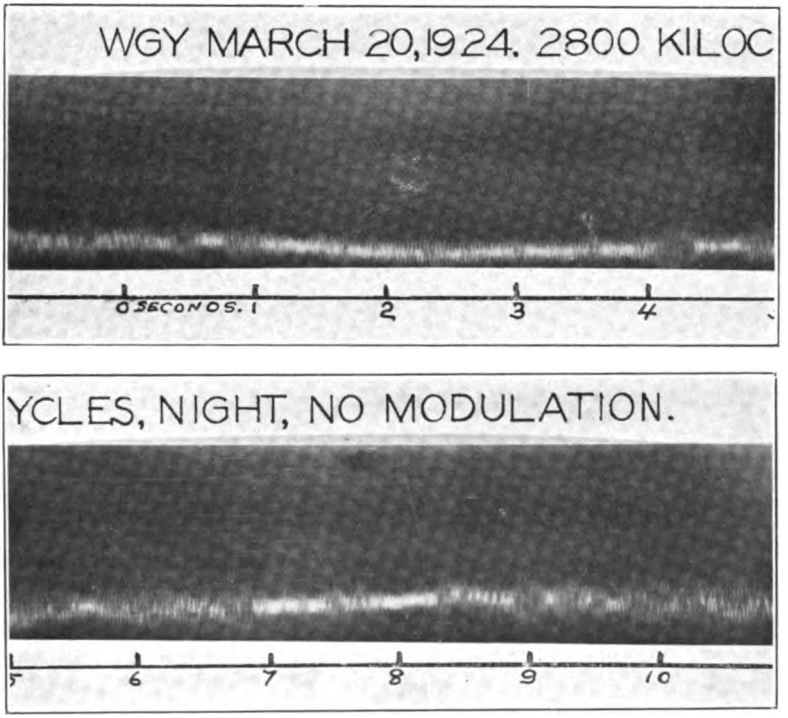

FIGURE 3

This record was made at a time when the station was not modulating, and shows a number of bursts of high frequency fading. Altho a careful examination of the record shows that

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these bursts are not exactly musical notes, but rather oscillograms of noise, frequencies of at least seventy cycles are not uncommon.

It is only necessary to listen in for an evening with fair amplification to observe that the frequency band from 3,000 to at least 6,000 kilocycles is far from a silent one. Altho the static and general noise background is usually several times less than in the normal broadcasting range, this region is harmonically haunted by many transmitters, including a number of broadcasters. At Newton I can usually hear WHN, New York, on its fourth harmonic at about 3,300 kilocycles; WFI, Philadelphia, on its fifth at 3,800, and even KYW, Chicago, on its eighth at 4,480, not to speak of the myriad amateur continuous wave harmonics. So far the high frequency band has been largely spared the wail of the mishandled regenerative receiver, that chorus of the damned which so sorely afflicts the normal broadcasting band. But now that the newspapers are publishing detailed descriptions of single circuit regenerators for 3,000-kilocycle reception, this happy state will soon vanish. Perhaps the only way any frequency channel can be swept free of this pest is to fill it with transmission of such a character that a very complicated reception system is needed for its unscrambling.

However, harmonic radiation from broadcasting stations is not entirely a waste product; I have found it distinctly useful. It is, for example, a most excellent thing with which to study high frequency reception at many other frequencies and distances than those of KDKA and WGY. In Figure 4 the upper record is from WEAN, Providence, Rhode Island, 60 kilometers (38 miles) distant, on its second octave at 4,400 kilocycles. This harmonic is very weak in Newton by day, but at night WEAN's field at this frequency is quite strong. The second record is from WBZ, Springfield, Mass., 118 kilometers (74 miles), also on its second octave at 3,560 kilocycles.

In the third record, shown for comparison, is WBZ at its fundamental of 890 kilocycles (337 m.), this record being made on the same evening and only a few minutes after the second record. It is interesting to note that reception from WBZ on the harmonic is generally far worse in quality than on the fundamental, as the high frequency is severely afflicted with ultra-rapid fading.

Altho based on very meagre data, my present feeling is that transmission in the neighborhood of 3,000 kilocycles is somewhat limited in its applications. It certainly does not carry speech or

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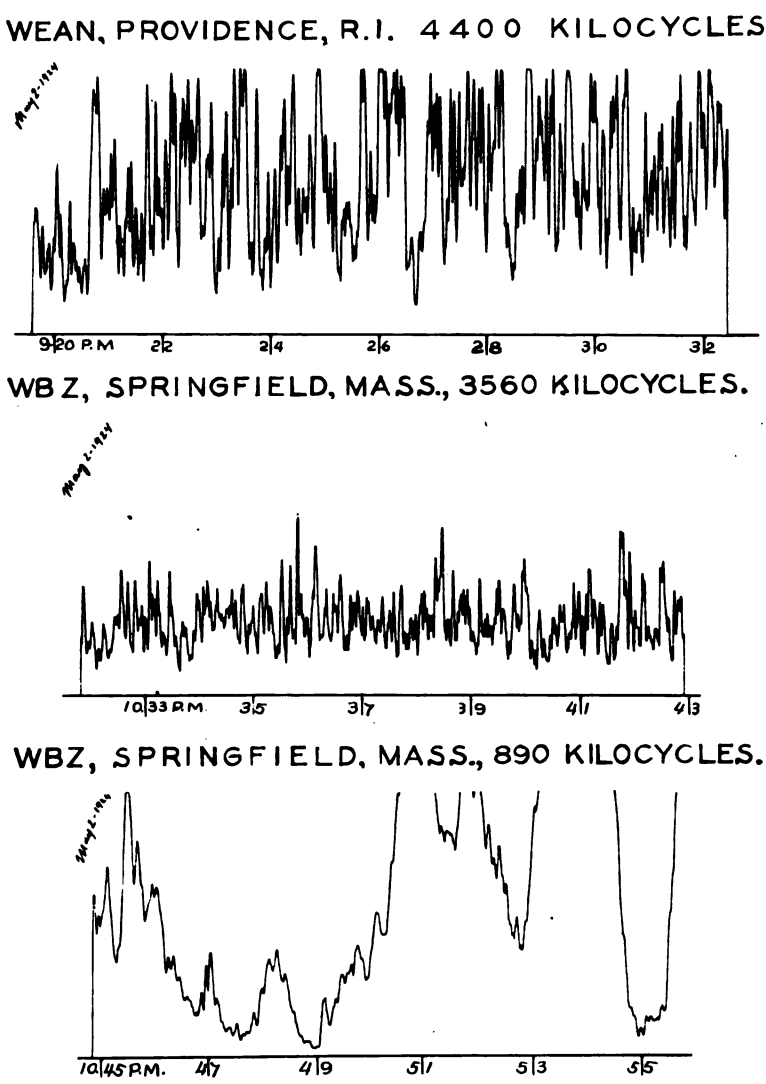

FIGURE 4

music well from Schenectady to Newton, a distance of 225 kilometers, and altho from Pittsburgh to Newton, a distance of 760 kilometers, the quality of the transmission is much better, it is still inferior to the normal 920-kilocycle radiation. Judging simply from my hotel experience, it may be an excellent means of reaching the city cliff-dweller in his cage of steel. Overland, for distances of over two or three hundred kilometers, its princi-

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pal usefulness is for night transmission, as the sun finally kills its field. Over salt water my observations of this frequency are limited to trans-Atlantic reception, and here, also, daylight is fatal. Simultaneous records made at slightly separated receiving points have shown such marked differences that some form of distributed pick-up is strongly indicated as a means of reducing distortion.

An extended investigation of transmission at frequencies materially higher than 3,000 kilocycles is now very much in order. Among other things, it would be interesting to compare the fading for a beam with that for distributed radiation; if our hypothesis have any real footing there should be a marked difference.

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BROADCAST TRANSMITTING STATIONS OF THE RADIO CORPORATION OF AMERICA*

By

Julius Weinberger

(Research Engineer, Technical and Test Department, Radio Corporation of America, New York)

The purpose of this paper is to describe the general arrangement, apparatus and operation of the radio broadcasting stations, with their associated wire line facilities, established by the Radio Corporation of America in the cities of New York and Washington, D. C., and placed in operation during the spring and summer of 1923.

The New York City station is known as "Broadcast Central," and transmits on two wave lengths or channels simultaneously, these being at present 455 and 405 meters (660 and 740 kilocycles). Its call letters are WJZ and WJY, respectively. The Washington station has a single channel only, with the call letters WRC, and transmits at present on 469 meters (640 kilocycles) wave length. All channels deliver approximately 500 watts to the antenna, at the carrier frequency. In all cases complete duplicate units of equipment are provided, with the exception of the antenna, and these can be thrown into action, in most cases, by means of relays, should failure occur in any part of the working equipment. This precaution has proven of great value in practice, in maintaining the uninterrupted service demanded of broadcasting stations located in large cities.

In addition to its own studio, each station has wire line facilities for the transmission of events occurring at outside points (for example, concert halls or theatres), these lines being connected to the radio transmitter thru appropriate means. Wire connections have also been established, linking the New York and Washington stations, as well as station WGY of the General Electric Company at Schenectady, New York, so that programs originating at any one of these stations may be transmitted from another, or simultaneously from all three.

*Received by the Editor, May 13, 1924. Presented before The Institute of Radio Engineers, New York, June 4, 1924.

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Broadcast Central

General

As mentioned above, this station was designed to transmit on two channels simultaneously. Originally the intention was to utilize one channel exclusively for classical music, instructional lectures, and serious types of entertainment generally, while the other channel was to be used for popular music, athletic events and lighter forms of entertainment; so that the listener might choose the particular kind of amusement which pleased him best. This plan has been adhered to, so far as possible.

In considering the location of the station, it was felt from the first that this should be conveniently accessible for artists, and should be such that short wire lines could be run directly to theatres and other places of amusement from which entertainment material might be obtainable.

The site finally chosen was the Aeolian Building in West 42nd Street, between Fifth and Sixth Avenues, New York City (Figure 1). This building is practically the musical center of New York, housing Aeolian Hall (where many concerts are given thruout the year), and the offices of the representatives of many of the foremost musicians and musical organizations. It is also near the center of the theatrical district, which makes it easily accessible for artists; and sufficiently far out of the highly attenuating tall building zone of lower Manhattan Island so that the signal intensity from the transmitting station was expected to be good in the residential sections of New York.

The Aeolian Company arranged to lease space to us on the sixth floor of their building for studios, control room, and offices, while the transmission apparatus was placed in a specially built house on the roof (about 200 feet (61.5 meters) above street level), where two self-supporting steel towers were also erected for the antennas.

The general arrangement of the station is shown in Figure 2 while a more detailed view of the arrangement on the studio floor is given by Figure 3.

On the sixth floor are located the studios for each channel, with suitable reception and coat rooms for artists. Here also are offices for the Program Division, and for the operating staff, which offices are not shown in the figure. Between the two studios, a room called the "Control Room" is placed (described in detail later on), from which the technical operation of the station is governed.

The furnishings and color scheme of the studios and their

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reception rooms were chosen with reference to the two-channel operation previously referred to. Thus the rooms of the "classical" channel, WJZ, were finished in dignified and somewhat restrained fashion, while those of the popular or "jazz" channel, WJY, were decorated in bright colors with suitable hangings. Photographs showing the interior of these studios are given in Figures 4 and 5, respectively. On the end wall of the WJZ studio a mural painting was placed (the work of Mr. Leon Soderston), depicting the many types of broadcast listeners. This

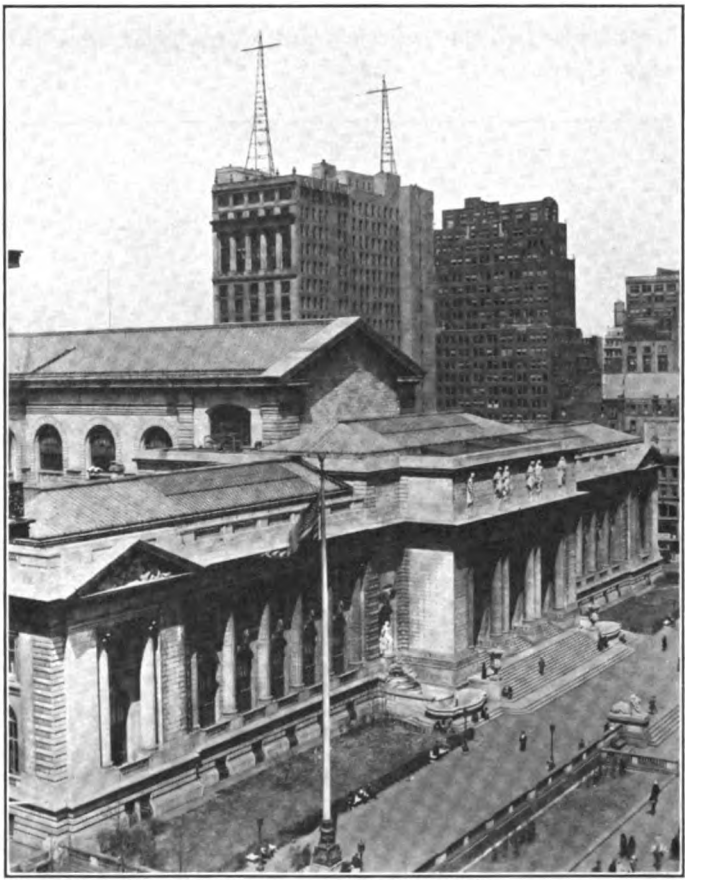

FIGURE 1—Site of Broadcast Central (Aeolian Building) New York

is faced directly by artists, when broadcasting, and by its appeal to the imagination helps to give them the feeling of a vast listening audience.

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The studio ceiling and walls are covered by an acoustical treatment[1] consisting of one inch (2.5 cm.) thick, special acoustical hair felt next to the wall, with a layer of muslin stretched over the hair felt and spaced from it by half an inch (1.25 cm.) of air space. The muslin is painted with a special paint which transmits sound, so that the felt may absorb it. When the treatment is finished the impression of a solid wall is given the casual observer. This acoustical treatment is carried down the walls to a point about 3 feet (0.92 meter) from the floor line, and from this height to the floor a solid material, such as gypsum wall board, is used. This is done in order to avoid the damage which might be caused to the stretched muslin covering, were it used, by careless persons.

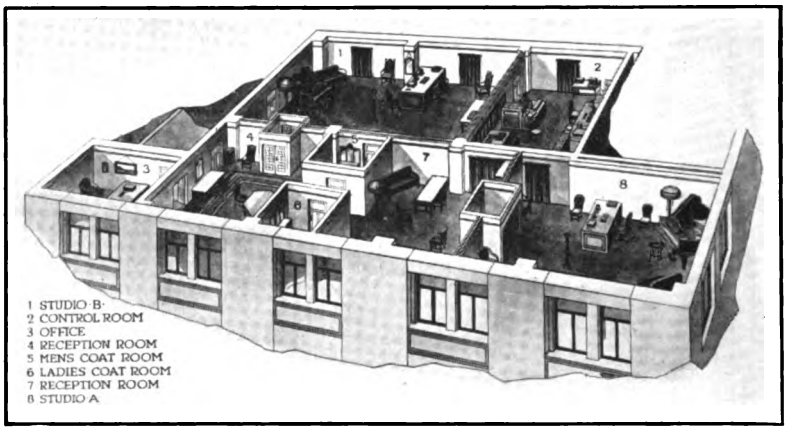

FIGURE 3—Arrangement of Rooms on the Sixth Floor, Aeolian Building

The studio floors are covered with thick carpet and the furniture is heavily cushioned. All of these acoustical measures are employed in order to reduce the "reverberation time"[2] of the studios to a minimum. This is a customary procedure in broadcasting studios and is frequently the source of question on the part of musicians as well as scientists.

W. C. Sabine found that good musical taste demanded a reverberation time of 1.1 second at 512 cycles for a room in which

[1] Installed by the Johns-Manville Company, New York City.

[2] See "Collected Papers on Acoustics," by Wallace Clement Sabine, Harvard University Press, Cambridge, Massachusetts. The reverberation time of a room is defined as follows: Assume a sound source to produce a sound intensity (energy), at a given point in a room, of 1,000,000 times the energy which would be just audible; if this source be stopped, the reverberation time is the time taken for the sound energy at that point to fall to a just audible value.

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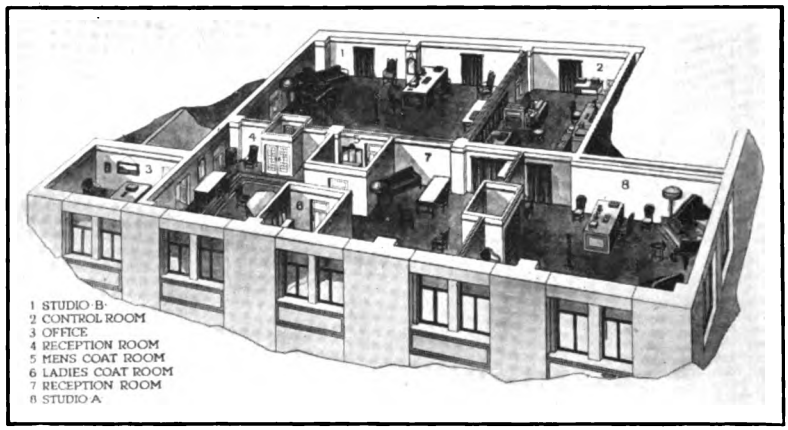

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piano music was to be given, in order to give the correct "ring" to the piano tones. However, were the studios designed for this value, objectionable resonance phenomena would be found, when using microphones to convert the sound to electrical energy.

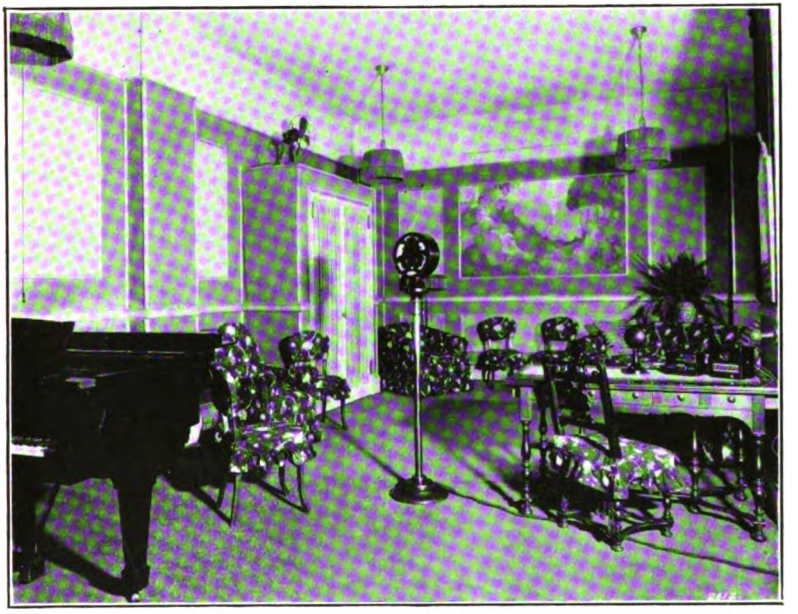

FIGURE 4—Studio of Station WJZ, Broadcast Central, New York

A possible explanation of this may be the following: When sound energy leaves a source, it spreads out in an air wave, the energy of which is partly contained in pressure variation and partly in the velocity of motion of the air particles. The wave energy is thus lodged in two components, in a similar manner to that of an electromagnetic wave, in which the energy is contained in an electric and a magnetic field. When the sound wave strikes an obstruction, such as the walls of a room, energy is partly transmitted, partly absorbed, and partly reflected, the proportion of each varying with the material and composition of the obstruction; so that at a given point in the room the resultant intensity at various frequencies due to sound coming directly from the source added to sound due to all reflections from obstructions, has one set of values for the pressure components and another set of values for the velocity components. Now, the human ear probably responds to both energy components in some measure, while all the microphone devices we have (with the exception

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of the thermophone) respond largely to the pressure component. Thus, when we place a microphone at a point within a room, we obtain a response indicating only the pressure fluctuations at that point, while with the ear both pressure and velocity fluctuations would be indicated. Thus if a very large pressure with small velocity existed at one frequency and a small pressure with large velocity at another frequency, the microphone would give a widely different response at the two frequencies, being operated mainly by pressure; while the ear, drawing its energy from both components, would give a fairly equal response.

FIGURE 5—Studio of Station WJY, Broadcast Central, New York

The usual solution of this difficulty is therefore to make the studios highly damped, so that there is as little reflection as possible from the walls. Then interference and resonance phenomena occur in a more limited measure, the microphone being arranged so as to be operated by the direct sound coming from the source. The calculated reverberation time of the WJZ studio is 0.4 second (at 512 cycles), as compared with the 1.1 second value considered correct for ear listening. It is likely, however, that as we learn how to make converters of sound to electrical energy which simulate the ear more closely in their re-

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sponse to the two energy components of sound waves, that the damping of studios will be reduced to a value more in accord with that for ear listening.

Returning now to the layout of Figure 2, it will be noted that the control room is connected by means of cables to the wire lines for transferring material picked up outside the studios, to the transmitter house on the roof, and to the Aeolian Concert Hall (Figure 6). Between the control room and the WJZ studio a long narrow window is placed, and a door opening into the WJY studio, thru which the control operators may observe the performance going on in the studios. The control and supervisory apparatus is located near these observation points.

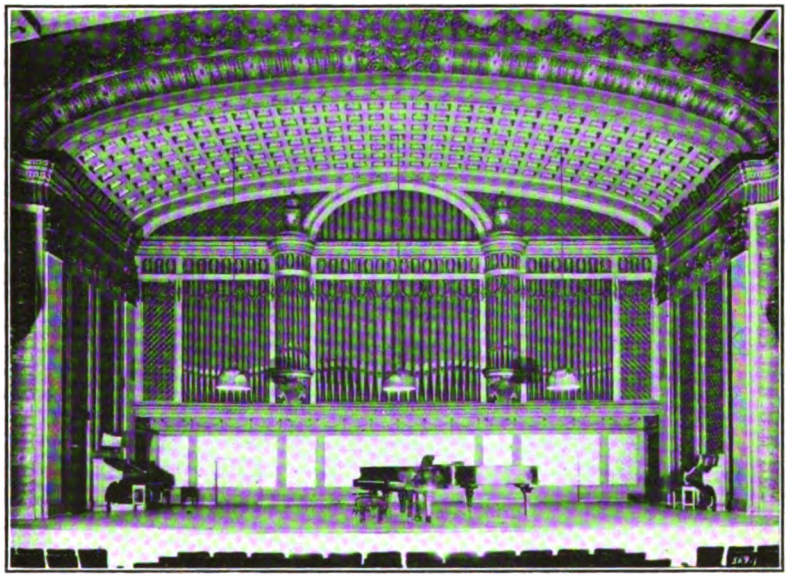

FIGURE 6—Stage of Aeolian Hall

A ventilating system is provided for blowing fresh air (heated in winter) into the studios and control room. An interesting point in connection with this is that by merely lining the ducts with hair felt 1 inch (2.5 cm.) thick, all noise from the ventilating motor is reduced to inaudibility in the rooms, and sounds emanating from one studio do not pass into the other, or into the control room, via the ventilating ducts.

On the roof, four transmitters (a working transmitter and a spare transmitter for each channel) and duplicate batteries for each channel are provided. The exterior of the transmitter house

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is shown in Figure 14 and the interior in Figures 15 and 16, while a view of the antenna towers[3] and the antennas is given in Figure 7.

FIGURE 7—Antenna Towers at Broadcast Central

STATION APPARATUS

The apparatus used in broadcasting may be divided into:

(a) Station apparatus, more or less fixed in place, used in connection with broadcasting from the local studios, or in connection with wire line transfer of material to be broadcast from points outside the studios.

(b) Portable apparatus, such as microphones and amplifiers, for picking up material outside the studios, and arranged to convert sound energy to electrical energy of suitable value for wire transmission to the station, there to be broadcast thru the station

[3] For the details of mechanical design of these towers as well as those of the Washington Station, see: Albert W. Buel: "The Development of a Standard Design for Self-Supporting Radio Towers," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 12, page 58, 1924.

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apparatus. The portable apparatus will be described in a subsequent section.

A schematic wiring diagram of the station apparatus involved in broadcasting on one channel is shown in Figure 8[4]. Starting at the upper left hand corner of the figure, we have single stage amplifiers which are arranged for inputs from a studio microphone or wire line (in case of broadcasting from points outside the studio). The microphone is connected directly to the input transformer, while in the case of wire lines a 40-mile resistance artificial line is connected between the line and input transformer to lower the input energy to a level of the same order of magnitude as that obtained from a microphone. The voltage produced by a microphone across the terminals of the input transformer is about two millivolts, while at the end of a line, a signal of about two-tenths of a volt is usually available.

The first stage is resistance coupled to the input side of a special amplifier designed to give uniform amplification over the frequency band from 100 to 4,000 cycles. From 4,000 to 10,000 cycles it is arranged to produce gradually increasing amplification. The reason for doing this is that all present loud speakers, head telephones, and many audio frequency transformers (as used in receiving sets) give a decreasing output in this range, and a rising frequency characteristic at the transmitting station will tend to compensate for this. As a result, such sounds as the consonant "s," the characteristic frequency of which is of the order of 10,000 cycles, are reproduced very clearly on the average receiving set, while music is also given greater clearness because of the better reproduction of harmonics.

The method of obtaining a flat curve for part of the range of this amplifier, and a rising curve for another part, will be evident from the figure. The output of the microphone amplifier is fed to the grids of two tubes, in parallel. The upper tube has a resistance in its plate circuit and the lower one a resonance transformer composed of air-core coils and a condenser with its resonance peak at 10,000 cycles. The voltage delivered by the upper (resistance) coupled tube is uniform with respect to frequency, while that of the lower tube is a resonance curve having a maximum at 10,000 cycles. The voltages delivered by these two tubes are added in series and applied to the grid of a third tube, the magnitudes being so chosen that the sum of the two gives a

[4]The station apparatus at Radio Corporation stations was originally built by the General Electric Company, and was modified in a number of details after some operating experience therewith. The description in this paper refers to the modified apparatus.

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frequency characteristic of the type mentioned above. The third tube serves as an output tube, and its output is stepped down in order to feed the input side of the power amplifier in the transmitter house thru a cable which is about 600 feet (185 meters) long. On the low voltage side of the output transformer is placed the main modulation control, a 1,000-ohm potentiometer, by means of which the voltage supplied to the power amplifier may be regulated. The transmitter house end of the cable is terminated by a 500-ohm resistance, from which the grid of the first power amplifier tube is fed. It may appear uneconomical not to place a step-up transformer at this point, but it is preferable to avoid transformers whenever sufficient amplification is already available, since each transformer cuts off transmission of low and high frequencies to some extent.

The power amplifier has two stages, the first being resistance coupled and the second transformer coupled to the grids of the modulator tubes. A step-down transformer with 500 ohms across the secondary is used so that with positive potentials applied to the modulator grids no distortion will occur, due to the grid current drawn. The customary "constant current" modulation system is used. The oscillators work into a local circuit (called the "tank circuit") which is inductively coupled to the antenna. Variometers are provided for wave length control in both local and antenna circuit. The antenna power, unmodulated, is normally 500 watts, antenna current about 7 amperes, on 455 meters and 6 amperes on 405 meters.

Supervision of the modulated antenna current is obtained in the control room by several methods (oscillographic observation of a small portion of rectified antenna energy, modulator grid current, and receiving set with loud speaker), the connections of which are evident from the diagram, and which will be discussed in more detail later on.

All units are supplied in duplicate, on each channel, but on the schematic diagram the various relays, jacks, plugs and switches for transferring from a working unit to a spare unit have been omitted, for clearness.

All filaments of vacuum tubes, except those in the transmitting set, are supplied from common 12-volt storage batteries in the transmitter house. The plate voltages of all tubes are supplied from the 2,000-volt motor-generator sets of the transmitters, the low plate voltages for the amplifiers being obtained from potentiometers fed thru resistance-capacity filters by one of the 1,000-volt commutators of these generators. The necessity of

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maintaining high voltage storage batteries for amplifier plate supply is thus eliminated, the supply from the generators being quite free from ripple after filtering.

For the sake of those readers who wish more details concerning the constants of the vacuum tubes used, reference is made to previous papers published in these PROCEEDINGS.[5],[6]

We will next consider some of the apparatus units in greater detail.

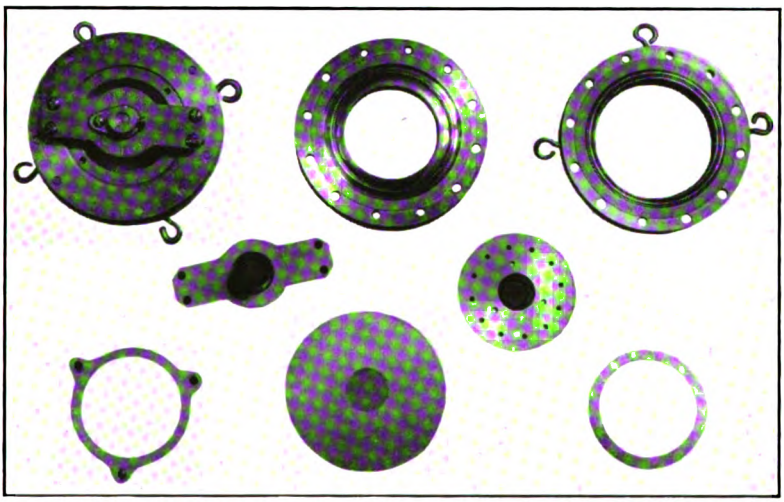

FIGURE 9—Double Button Carbon Microphone Assembled and Disassembled

STUDIO APPARATUS

The microphones used are modelled after the double button carbon type originated by the Western Electric Company.[7] This transmitter contains a stretched diaphragm of duralumin, about 0.0015 to 0.002 inch (0.0038 to 0.005 cm.) thick, spaced 0.001 or 0.002 inch from a metal surface, on one side, so as to secure high air damping;[8] carbon containing chambers are placed one on each side of the diaphragm, and current supplied to the microphone

[5] W. R. G. Baker: "Commercial Radio Tube Transmitters." PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 11, page 601, 1923 (particularly the tables on pages 653 and 654).

[6] J. C. Warner: "Recent Developments in High Vacuum Receiving Tubes —Radiotrons Model UV-199 and Model UV-201-A," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 11, page 587, 1923.

[7] I. W. Green and J. P. Maxfield: "Public Address Systems" (Section entitled "Transmitters"), "Journal of the American Institute of Electrical Engineers," April, 1923.

[8] I. B. Crandall: "The Air Damped Vibrating System," "Physical Review" 2nd Series, volume XI, page 449, 1918.

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as shown in the schematic diagram of Figure 8. Normally the maximum permissible current thru each button is 20 milliamperes, and with this each button has an internal resistance of about 100 ohms; thus the transmitter forms a generator (as applied to the input transformer) having an internal resistance of about 200 ohms. The voltage produced across a 200-ohm load is about 5 millivolts for one dyne per sq. cm of sound pressure in air.

The diaphragm characteristics are such (due to its high natural frequency and great damping), that practically uniform output is secured, for equal incident sound pressures over the frequency range from 30 to 7,000 cycles. Higher frequencies up to 10,000 cycles are usually reproduced well with fresh carbon, and when the transmitter current is first switched on, but after some running, the reproduction of these frequencies falls off. However, for nearly all work, these transmitters are quite as satisfactory in their reproduction as a condenser transmitter, and are simple, require much less amplification than the condenser transmitter, are conveniently portable, and may have as long leads as desired attached to them (which can be done only in a limited degree with the condenser transmitter). Figure 9 is a photograph of one of these microphones.

In the studio, two microphones are hung on spring suspensions in the stand shown in Figure 10, for picking up studio concerts, one microphone being in operation and the other a spare capable of being switched in from the control room to replace the working microphone.

For announcing, a single microphone only is usually found sufficient, altho provision was originally made to have a spare for this purpose also. This is suspended in a small support of convenient height.

A control box (Figure 11) is also placed on the announcer's table. This contains a switching key, by means of which the announcer may connect his own microphone, or the studio microphone to the microphone amplifier ~~in the~~ control room, or cut off both. This key operates relays in the control room which switch on filament current to the single stage amplifier connected to the particular microphone which is operating. A small signal lamp, mounted above the switching key, remains lit automatically while current flows in the antenna, so that the announcer is thus informed, should the antenna power go off at any time. The signal current is supplied to this lamp by means of a special relay in the transmitting set, connected in series with the oscil-

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lator grid leak, so that it is operated by the direct current flowing thru the leak.

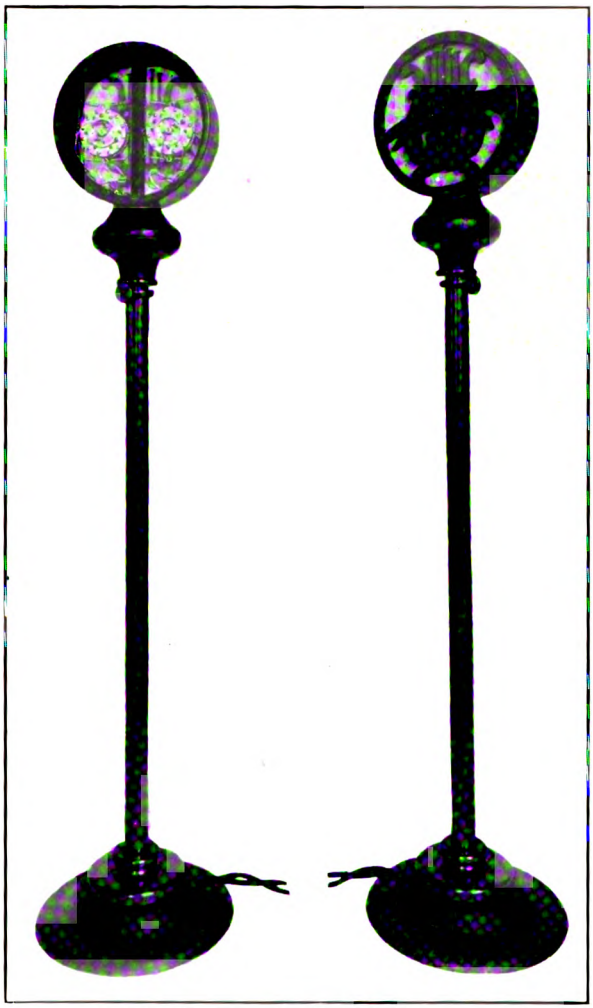

FIGURE 10—Studio Microphone Stand, Open and Closed

CONTROL ROOM APPARATUS

The apparatus here is the following:

1. A single stage amplifier rack, shown in Figures 12 and 13. This contains, in the upper row, two amplifiers suitable for microphone input and one for input from telephone lines, and in the second row, duplicates of these. Any amplifier may be connected

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FIGURE 11—Announcer's Control Box

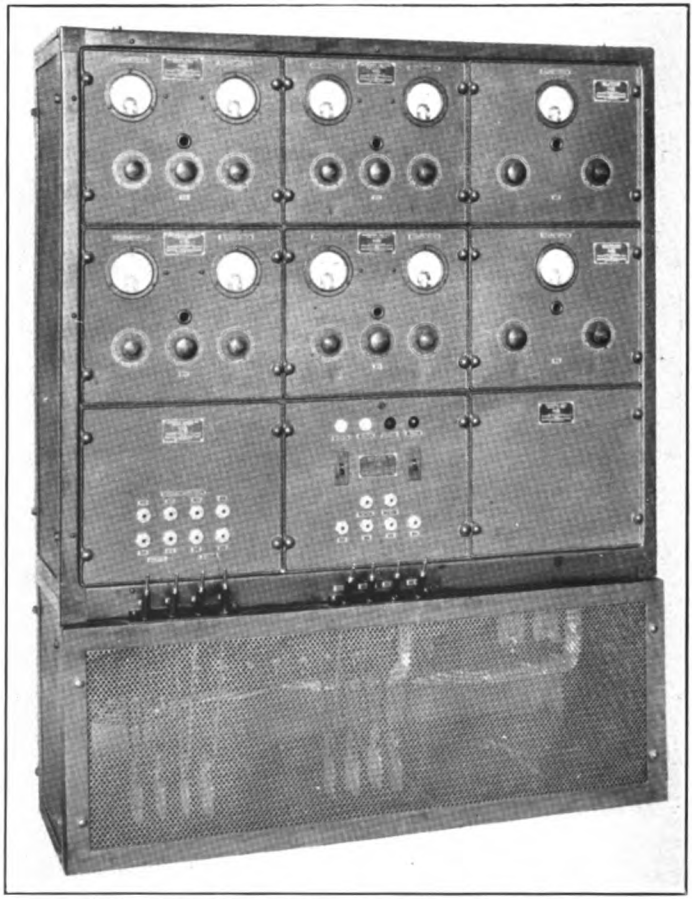

FIGURE 12—Control Room Amplifier Rack, Front

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to any microphone by means of the jacks and plugs shown in the lower row of panels. Here also are jacks for listening to the output in the plate circuit of whichever single stage amplifier happens to be in operation, and also for listening to the radio output (conducted thereto from the monitoring receiving set). Signaling lamps are also provided, one to indicate when either concert or announce microphone is in operation in the studio, one to indicate the presence of antenna power, and two for signaling between control room and transmitter house, if desired. Each single stage amplifier is equipped with filament voltage control and voltmeter, microphone rheostat, ammeter (in microphone amplifiers), and an input intensity control. The latter is a small carbon plate compression type of rheostat, which is quite noiseless in operation, permitting a shunt resistance of from about 50 ohms upward to be placed across the input transformer. All tubes are the UV-202 type, rated at 5 watts as oscillators, having tungsten filaments.

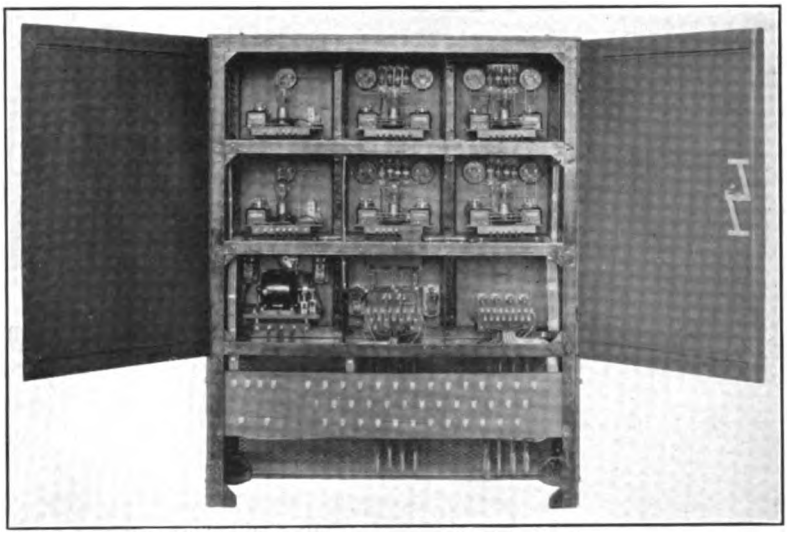

FIGURE 13—Control Room Amplifier Rack, Rear

This amplifier rack also contains relays by means of which the filament circuits of the amplifiers which the control operator has selected for the announcing or concert microphones are closed when the key in the announcer's control box is operated.

2. A special amplifier for producing a rising gain in the frequency range from 4,000 to 10,000 cycles; this is mounted above the microphone amplifier, as shown in Figure 14, duplicate equip-

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ments being provided. Its wiring was indicated in the schematic diagram of Figure 8, and its operation has already been described. In the operation of the station it is usually referred to as the "S amplifier" because of the marked improvement it produces in the reproduction of this usually difficult consonant.

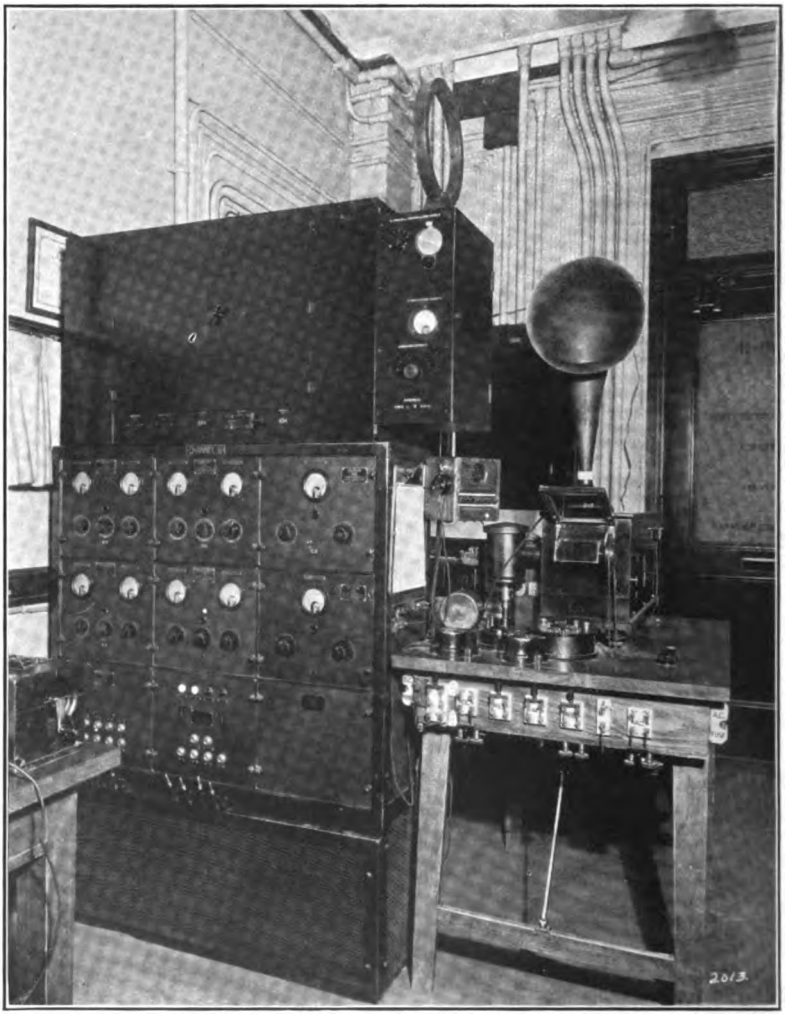

FIGURE 14—Control Room Amplifier, "S Amplifier," Modulation Control and Supervisory Equipment

3. A monitoring receiver consisting of a small loop, tuning condenser, detector, and two stage audio frequency amplifier. This receiver has approximately the same audio frequency repro-

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duction characteristics as the average good present-day receiver, so that what the control operator hears has the same quality as would be heard by the average listener. The receiver is in the box to the right of the "high audio frequency" amplifier, in Figure 14. In the same box are placed the main modulation control, and a galvanometer which indicates the grid current taken by the modulator tubes in the transmitting set. The latter is used for supervisory purposes.

4. A General Electric Company oscillograph, one vibrator of which is fed from rectified antenna current. (See diagram, Figure 8). This appears below the monitoring receiver. Lines are marked on the ground glass screen to indicate full antenna current and fractions thereof, up to double and down to zero. These lines and the vibrator spot are reflected in rotating mirror, and the control operator thus has a visual picture of the modulated antenna current.

The subject of supervising will be treated further in a subsequent section.

The connections between control room and transmitter house, for amplifier power supply, audio frequency, rectified modulated radio frequency, and signaling, are all lead-covered cable in iron conduit; this method of connection is quite adequate for keeping radio frequency out of the audio frequency lines. The conduits are run thru the hall to the rear elevator shaft of the building and then up thru this to the roof, a total distance of some 600 feet (185 meters).

The control room also contains all equipment for wire line transfer of incoming or outgoing material for broadcasting; this will be described in another section.

Transmitter House

An external view of this is shown in Figure 15. It is a building 37 feet (11.4 m.) long by 20 feet (6.2 m.) wide by 12 feet (3.7 m.) high, inside.

Internal views are shown in Figures 16 and 17. Figure 16 shows a view of the following: at the extreme front, left, the power amplifier of one channel, beyond this the operating tables, and beyond this, a power amplifier of the second channel; beyond this last, the power switchboard for the direct current supply to the motor-generator sets. At the right side of the picture are four transmitters, with their motor generator sets behind them. The high panels are the transmitting sets, and between each pair of transmitters is a panel containing the filters and potentiometers

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for plate power supply to the various station amplifiers, derived from the transmitting motor generator sets. At the distant end of the room can be seen the oscillograph rectifier and coupling to the antenna.

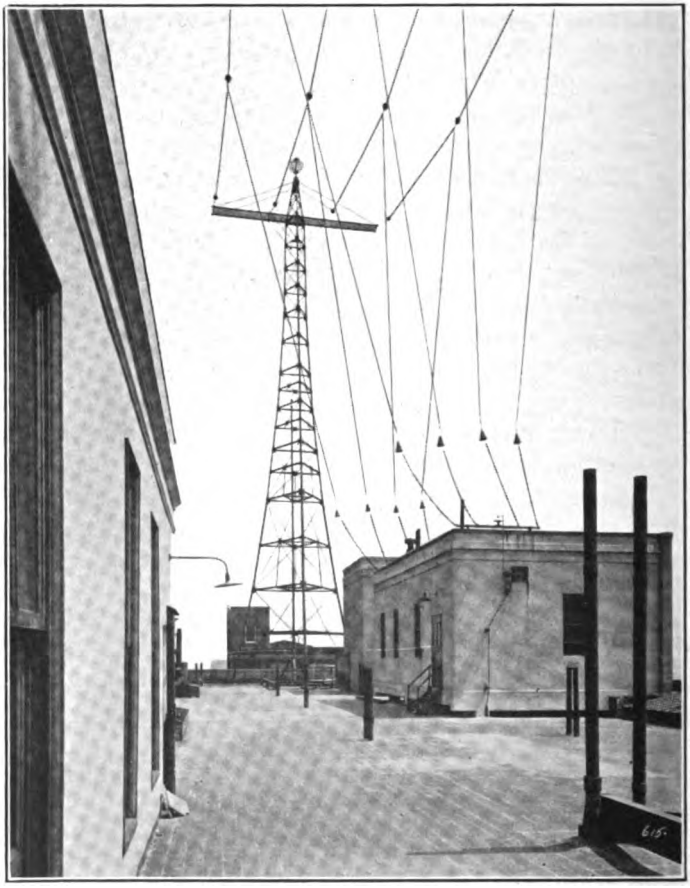

FIGURE 15—Transmitter House, Broadcast Central, Exterior

Figure 17 shows a rear view of the transmitting sets and motor generators. One motor generator set is associated with each transmitter, and bears on a common bedplate the driving motor, a combination 110-volt exciter and 30-cycle volt alternating current generator for the filament power of the transmitters, and a 2,000-volt direct current generator for plate power. The last-named has two commutators, 1,000 volts each, and from one of these plate power for the amplifiers is taken to the filter panel.

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FIGURE 16—Transmitter House, Broadcast Central, Interior

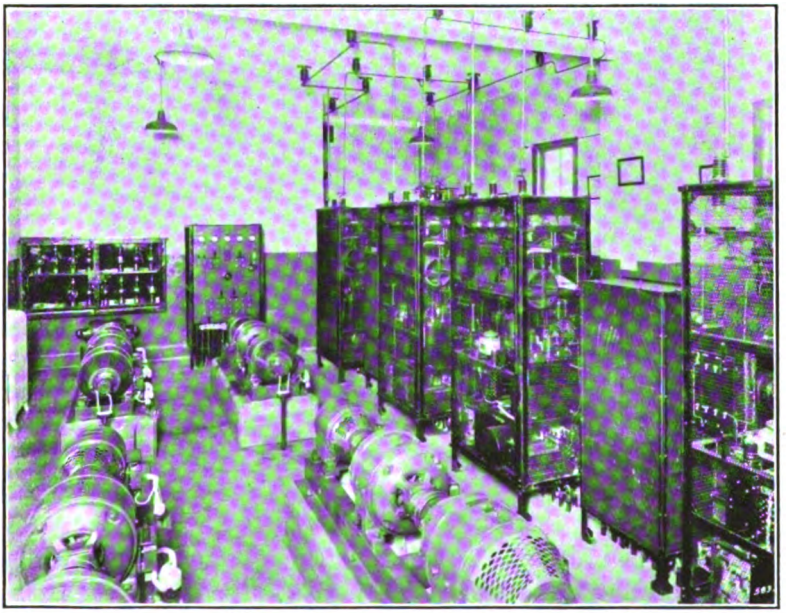

FIGURE 17—Transmitter House, Broadcast Central, Interior

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At the end wall are the storage battery charging panel and charging motor-generator, as well as a rack for holding transmitting tubes.

The interconnection of the various units will be evident from the schematic wiring diagram of Figure 8.

A closer view of a power amplifier rack is given by Figures 18 and 19. Here three stages of amplification are provided, of which normally only two stages are used. The number of stages

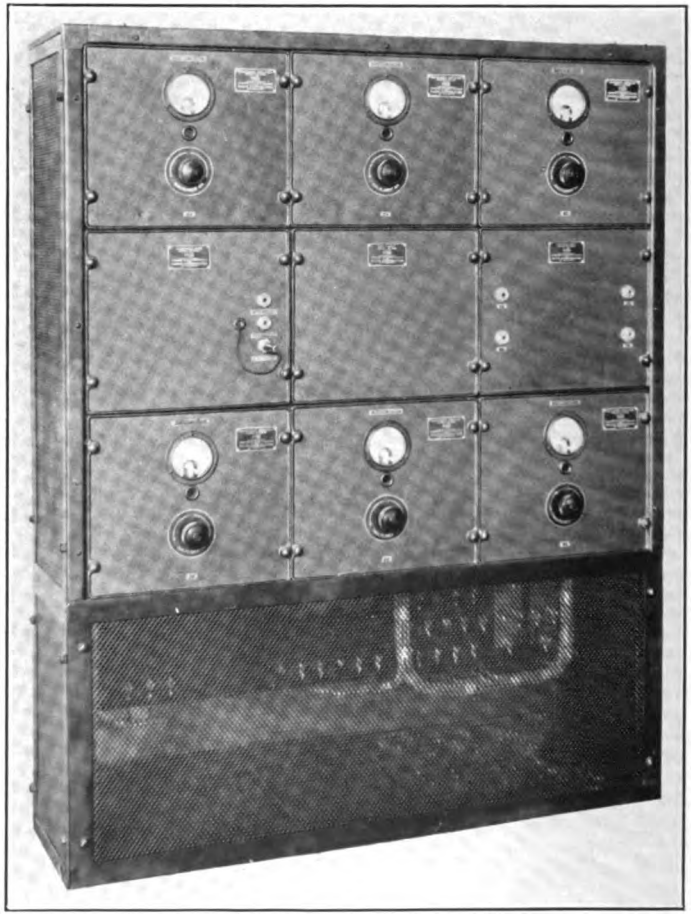

FIGURE 18—Power Amplifier, Front

desired can be connected by the plug and jacks on the left hand center panel. All tubes are of the UV-203-A type, except the last, which are UV-211, rated at 50 watts when used as oscillators,

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and having thoriated tungsten filaments. The amplifier stages are placed in the upper and lower rows of panels (working and spare units), the center row containing relays by means of which either amplifier can be cut in, these relays being controlled by change-over switches in the control box on the operator's table. Filament voltmeters and voltage control are the only controls provided here. On the right-hand panel in the center row (Figure 18) are listening jacks for checking the amplification in each stage. Each jack is shunted by a 500-ohm resistance, and this is placed in series with a fairly high resistance (of the order of hundreds of thousands of ohms) in parallel with the grid leaks of the respective amplifier tubes. The series resistances are adjusted to give equal loudness at each listening jack.

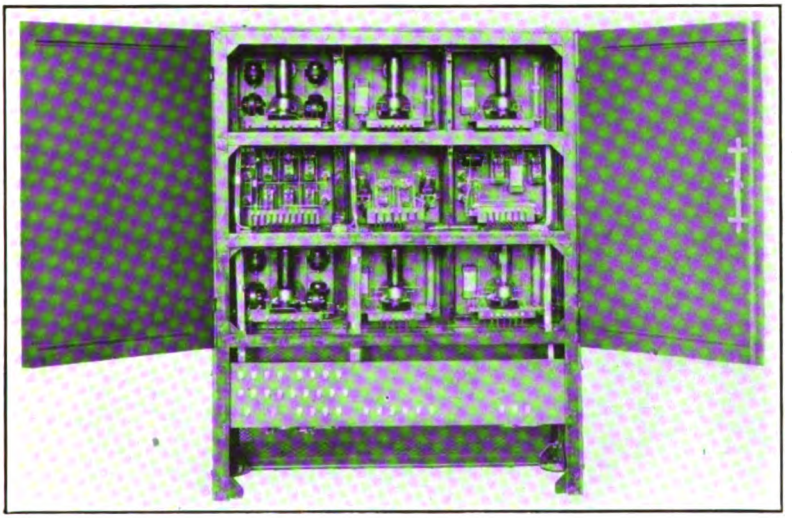

FIGURE 19—Power Amplifier, Rear

The connection from the output of the power amplifier to modulator grids is made by wire shielded by means of grounded copper tubing run overhead. These are the leads coming from insulators, on top of the power amplifiers in Figure 16. Shielding was found essential in order to prevent radio frequency currents from the antenna leads being induced in this circuit and causing the modulator tubes to overheat because of the high voltages impressed on the grids; in addition to the shielding, a small condenser (0.005 microfarad) is placed between modulator grids and filaments, in the transmitting set.

One of the transmitter panels is shown in Figures 20 and 21.

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The schematic circuit of these is indicated in Figure 8, and it will be noted that two oscillator and two modulator tubes are employed. Each tube is of the UV-204-A type, rated at 250 watts as an oscillator, and having thoriated tungsten filaments. On the

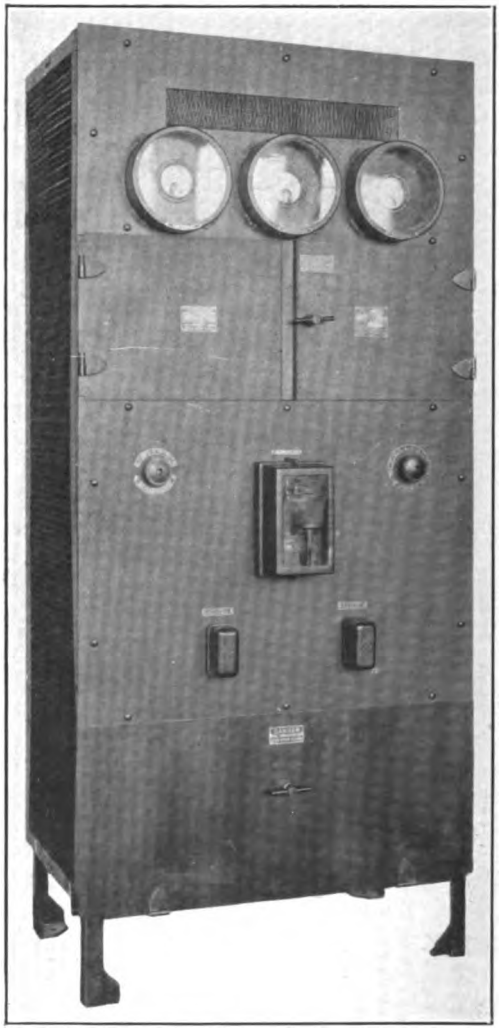

FIGURE 20—Transmitter Panel, Front

front panel are, respectively, tank circuit and antenna ammeters, plate supply voltmeter, controls for filament and plate voltages, voltage regulator (to maintain constant filament voltage if motor-generator speed varies), an overload relay (which shuts down

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the motor generator set if plate current becomes excessive thru poor circuit adjustments or tube failure), and the relay which lights the antenna power signal lamps in the control room and studio. The rear view of the transmitter (Figure 21) shows at the bottom, left, a multiple contact automatic switch (operated by push button from the operator's control box), which connects the antenna and performs various other functions to put the set in operation, and an automatic motor starter; beside these is the filament power transformer, and at the right the smoothing con-

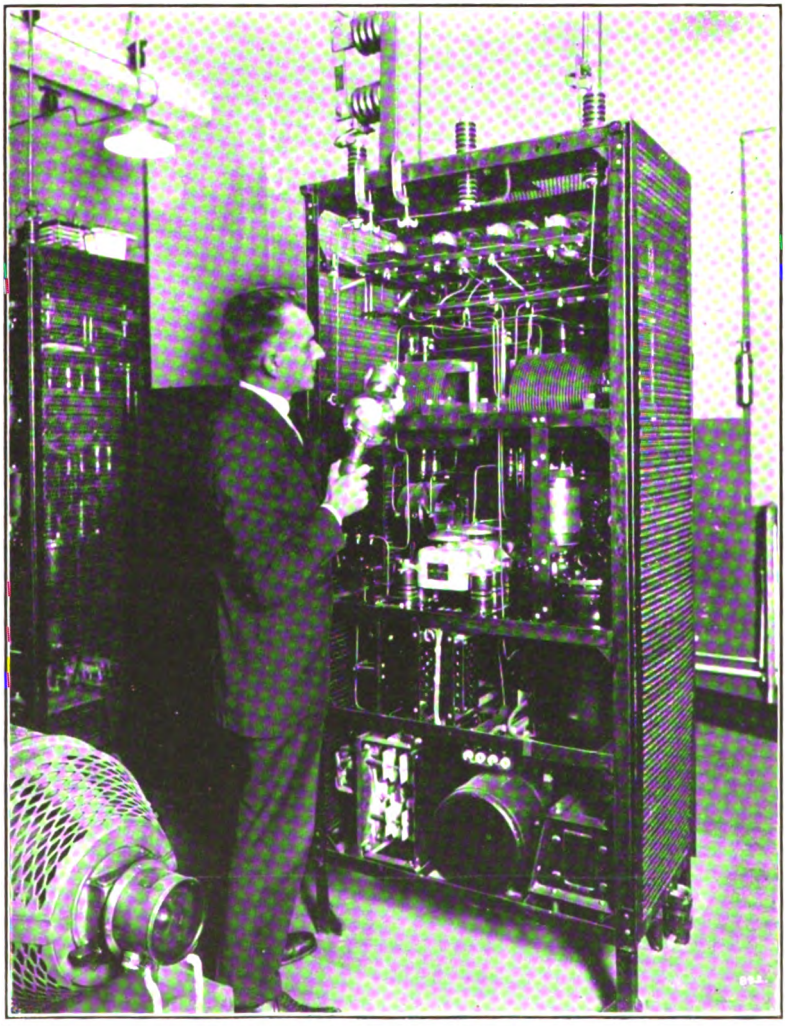

FIGURE 21—Transmitter Panel, Rear

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densers for the plate power supply. Above these condensers is the plate reactor, to the left of it are various condensers used in the oscillating circuit. Immediately above these are the tank circuit and antenna tuning coils, of edgewise wound flat copper strip, and at the top of the set are the transmitting tubes in a spring-suspended cradle.

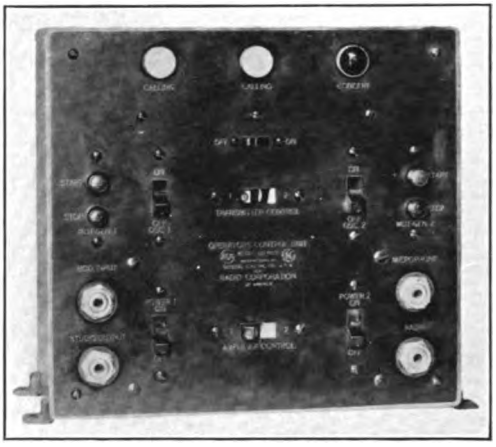

FIGURE 22—Transmitter Operator's Control Box, Front

On the operator's table is placed a control box, for each channel, which is shown in Figures 22 and 23. In this, viewed from the front, are placed the following controls for relays in the various units: near the bottom, filament power switches and change-over switch for either power amplifier, and at the sides, monitoring jacks for the input and output of the power amplifier and the radio output. In the middle at the sides are placed push-button switches for starting either motor-generator set, for throwing on plate and filament power, and for connecting either transmitting set to the antenna. At the top of the box are signaling lamps to the control room, and a lamp indicating that the announce or concert studio microphone is in operation. Thus in a very short time any distortion or trouble can be located by means of the monitoring jacks as arising in some particular section of the equipment, and a spare substituted for it.

A receiving set is also located on the operating table, and here a continuous watch is maintained so that the station may cease operation at once in case of marine distress signals; this is legally required of all coastal radio stations. The same receiver is utilized for the reception of time signals on a wave length of 2,500 meters

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(120 kc.), from Arlington (distance 220 miles) (350 kilometers), and these are re-transmitted on the 455-meter (660 kc.) wave by connecting the output of the receiver's audio frequency amplifier to the input of the transmitting power amplifier thru a suitable transformer. No difficulty is experienced in receiving on wave lengths of 600 meters (500 kc.) and above, due to interference by the local transmitters; the receiving antenna is run away as far as possible from the transmitting antennas, a coupled trap is connected across the antenna and ground posts of the receiver, this trap being tuned to the wave lengths to be received, and all equipment is housed in a metal case.

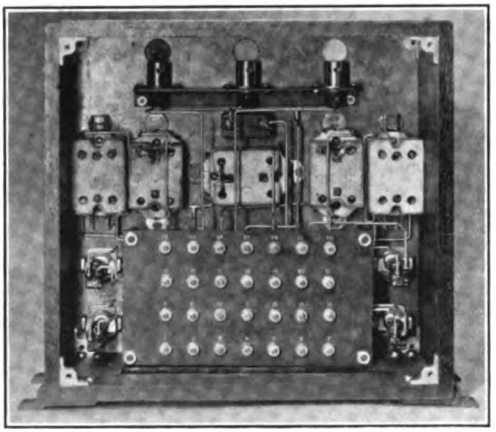

FIGURE 23—Transmitter Operator's Control Box, Rear

Figure 24 shows in somewhat better detail the following: at the bottom, one of the oscillograph rectifiers (containing a 50-watt, UV-203-A tube with grid and plate connected in parallel so as to function as a rectifier); above this, the coupling arrangement of the rectifier to the antenna, consisting of a single turn of the antenna lead placed within a helix; above this, a trap circuit necessitated by duplex operation, of which more will be said in a subsequent section; and finally, the antenna leading-out insulator.

The antenna system is shown in Figure 25, each channel being connected to an inverted "L."

SUPERVISION DURING OPERATION

In the description of control room apparatus the following were mentioned as being furnished for supervisory purposes to the control operator:

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(a) A receiving set, containing a vacuum tube detector, and two stage audio frequency amplifier, with a loud speaker, to represent the average listener's receiver.

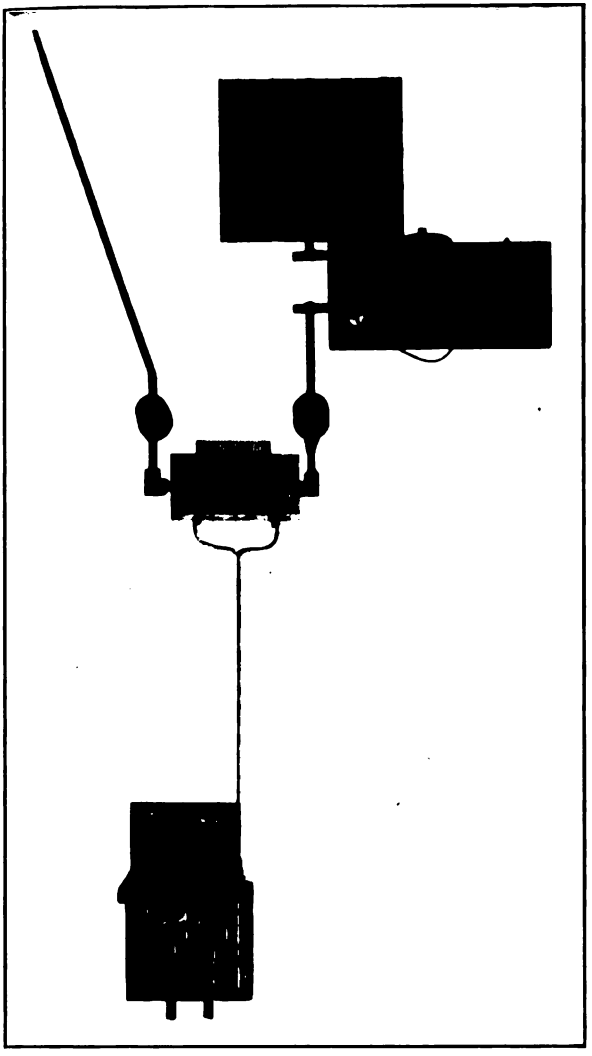

FIGURE 24—Oscillograph Rectifier with Antenna Coupling Arrangement, and Wave Trap

(b) An oscillograph, to one vibrator of which a small amount of rectified antenna current is brought, the fluctuations of the vibrator spot being made visible in a rotating mirror.

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(c) A galvanometer which indicates the direct current drawn by the grid circuit of the modulator tubes, when positive voltages are impressed.

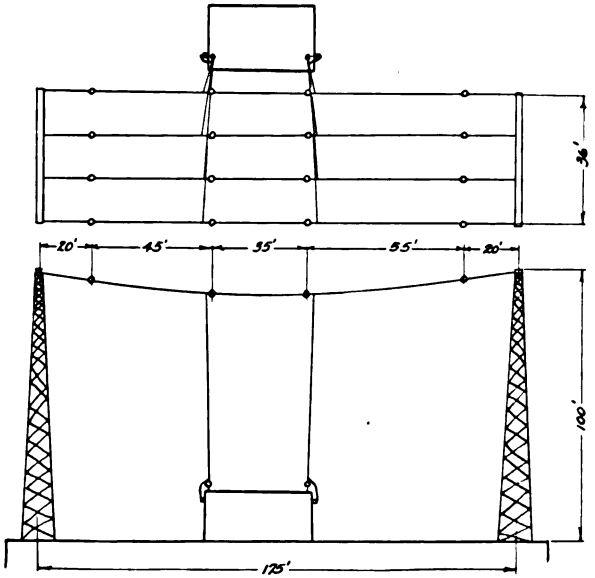

FIGURE 25—Broadcast Central, Antennas for Two-Channel Transmission

The control operator is concerned first, with maintaining as high a percentage of modulation (that is, ratio of side bands to carrier wave) as possible, without distortion; and, second, with the determination of the proper placing of artists in the studio, so that correct balance is obtained between soloists and accompanists, or between the various instruments of an orchestra. The receiving set is used to determine the general character of the performance, the presence of distortion, or to indicate any sudden interruption. The oscillograph indicates the percentage of modulation, while the modulator grid current galvanometer is an indicator which warns the operator that the maximum permissible percentage of distortion-free modulation is being obtained. The manner in which the last-named instrument is used is as follows: observation of the shape of the modulated wave in the oscillograph mirror will indicate that, in general, it is very irregular, having occasioned high "peaks" and many smaller irregular undulations; and it is essential to regulate the degree of modulation so that these peaks will not exceed a permissible

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maximum. Now, in the oscillograph, these peaks occur very rapidly, and are intermixed with all the other irregular fluctuations, so that for continual observation the eye would soon tire of judging the percentage modulation this way. However, it is really not necessary for the control operator to know the actual percentage at all times; all he is concerned with is that the modulation shall not *exceed* a certain maximum. This is conveniently done by the use of the modulator grid current indicator. Normally, with modulation less than about 30 percent (a fluctuation of antenna current from 30 percent below maximum to 30 percent above), no positive audio frequency voltages are impressed on the modulator grids, hence no grid current flows, and the galvanometer remains at zero. Whenever the modulation peaks exceed this value, a rapid "kick" of the galvanometer will be seen at each of the occasional peaks; thus this instrument is really a peak-reading ammeter. The presence and magnitude of these occasional deflections warn the control operator that modulation in excess of 30 percent is being obtained; normally a modulation of about 40 percent is allowed, corresponding to a certain deflection of the galvanometer. A greater degree than this would result in the generation of double frequency harmonics of appreciable magnitude, in receiving sets using square-law rectifying detectors, when sinusoidally modulating the transmitting set at a given audio frequency.[9] Thus the control operator need only watch to see that the occasional kicks of the galvanometer pointer do not exceed a certain value; most of the time the pointer will remain at zero.

This sort of observation method is considerably better for control purposes than one which involves the watching of a continually flickering object or wave, altho observation of the oscillograph is useful in making more exact observations of the character of the modulating wave and as a check against the galvanometer.

Audio Frequency Characteristics of Station Apparatus

In a broadcasting system it is necessary to transmit a range of frequencies, from the source of sound to the ear of the auditor, which up to the present we have taken as lying between 30 and 10,000 cycles. The frequencies which occur in music, in fundamental tones, range from 16 cycles for the largest organ pipes up to about 4,000 cycles for the highest instrument tones; how-

[9] R. V. L. Hartley: "Relations of Carrier and Side Bands in Radio Transmission." Proceedings of The Institute of Radio Engineers, volume 11, page 34, 1923.

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ever, harmonics which characterize the different instruments, or the singing voice, require the effective transmission of frequencies up to at least 10,000 cycles, in order to secure the clear, "keen" quality appreciated by musicians. Another aspect of these transmission requirements is that we find, experimentally, with a transmission range extending as high as 10,000 cycles, that each instrument or group of instruments in an orchestra seems to stand out clearly when reproduced in a good receiver, while limitation of the upper frequency transmission to 3,000 cycles, for example, causes the musical rendition to become "muddy" or blurred. The latter effect probably occurs because with a combination of instruments very rapid transients or changes in the sound wave shape may occur, taking place in a time much shorter than that corresponding to the fundamental frequencies being played; but a transmission system which will transmit frequencies up to 10,000 cycles will also transmit sudden fluctuations in wave shape taking place in a ten-thousandth of a second.

In the case of speech, the transmission of such consonants as "S" requires effective transmission of a frequency of the order of 10,000 cycles; similarly, to secure natural voice quality, particularly of male voices, it is necessary to transmit effectively frequencies down to perhaps 30 cycles.

The foregoing statements are mainly the result of qualitative observations on apparatus of which the frequency range of transmission was known, when listening with ordinary telephone receivers (as the broadcast listener does). We have not had any opportunity to determine in detail what the permissible change may be in the ratio of amplitudes of one frequency to another, as it exists in the initial sound, compared to what it may become by the time it reaches the listener, before the listener is aware of distortion.

The extent to which we have been able to meet the foregoing transmission requirements, in the station apparatus, is shown in Figure 26, and the method of taking the observations in Figure 27. This curve represents the amplitude of the alternating voltage applied by the modulators to the plates of the oscillator tubes (to which the modulation will be proportional), when a constant sinusoidal voltage of variable frequency is impressed in place of the microphone (a resistance of 200 ohms being placed in series with the voltage source to represent the internal resistance of the microphone). The microphone itself may be considered as sensibly such a source. The reason for the rising gain

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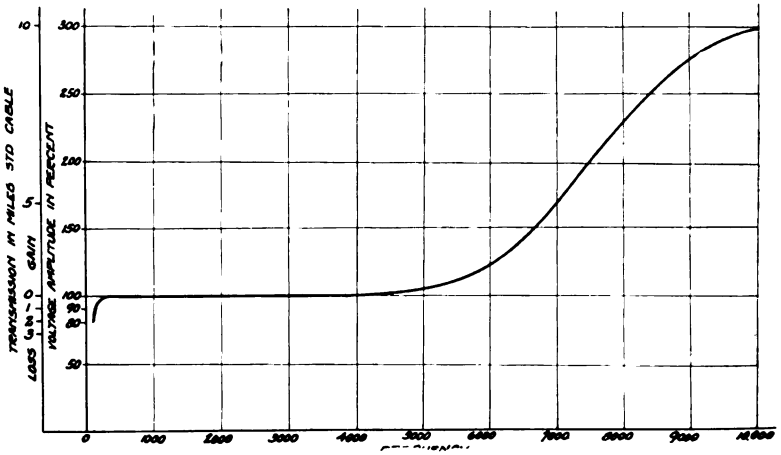

FIGURE 26—Performance Curve of Station Apparatus

at the high frequency end of the range has been mentioned previously.

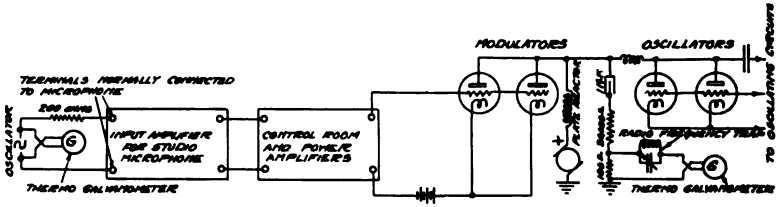

FIGURE 27—Method of Measuring Audio Frequency Modulation Characteristics of a Radio-Telephone Transmitter

TWO-CHANNEL TRANSMISSION

The operation of two channels simultaneously, at Broadcast Central, was attended at the start with difficulties due to "cross-talk" between the two channels. This cross-talk was due to modulated radio frequency currents induced from the transmitting apparatus or antenna of one channel in circuits connected with the other channel; these currents were rectified by vacuum tubes or other apparatus capable of rectifying action, became audio frequency currents, then passed thru the audio frequency amplifiers to the modulators, and modulated the transmitting set. Thus the audio frequencies modulating each channel might be heard on the wave length of the other channel.

No pure audio frequency cross-talk between the two channels was experienced, since audio frequency circuits where cross-talk might have been set up were only the connections between trans-

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mitter house and control room, and here each channel was run in lead-covered cable, in separate iron conduits.

Radio frequency induction was found in the following circuits, and eliminated in each case by the measures indicated:

(1) In studios and control room, the microphone leads to the amplifiers required very careful shielding, copper sleeving over the flexible microphone conductors being sufficient. The shielding was continuous from the amplifier terminals to the microphones; in a first installation, the wiring had been run from control room to studio, thru lead-covered cable between terminal boards located in these rooms, in standard electrical fashion, and in each terminal box a few feet of unshielded cable had been left. These few feet were quite sufficient to give rise to serious cross-talk.

(2) The control room amplifiers themselves, being in steel cases, picked up no appreciable amount of radio frequency. However, there was always a considerable amount of radio frequency which came in on the outside wire lines, when these were connected to the amplifiers. This was prevented from reaching the first amplifier tubes by the use of audio frequency input transformers having a copper shield between primary and secondary windings. In some cases cross-talk was found even after this was done; this was traced to the rectifying action of some of the carbon block compression rheostats, which were shunted across the primary windings of all input transformers, for modulation control purposes, in the original installation. This control method was superseded later on by the arrangement shown in Figure 8.

(3) In the transmitter house, the proximity of the two antennas caused a current of 1 to 2 amperes from each channel to flow in the down-lead of the other channel. This current eventually caused radio frequency voltages to be induced in the grid circuit of the modulator tubes, large enough to give serious cross-talk. The difficulty was remedied by inserting coupled trap circuits in each antenna lead, as shown in Figure 28. Each trap was tuned to the wave length of the opposing channel, and the coupling tap adjusted to display as high an effective resistance as possible at the wave length of the opposing channel, with still no appreciable loss of energy at the wave length of its own channel. The method of adjustment was simply to note the current in the antenna lead with each channel running alone, and adjust so that the cross-talk current was made as small as possible, without affecting the desired antenna current; all circuits were left com-

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plete thruout, as for two channel operation, and oscillations started by applying plate voltage.

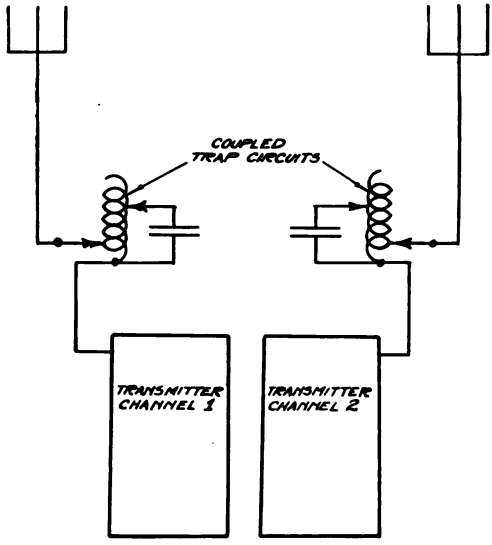

FIGURE 28—Trap Circuits for Reduction of Cross Talk Between Channels

(4) When using portable amplifiers on work outside the studio within a distance of about a mile from the station, or of other radio stations, and in certain locations, considerable radio frequency was found present in the wire lines to the station. These lines are often overhead twisted pairs, which act as a very good antenna. The radio frequency current flowing in these lines would cause voltages to be induced in the apparatus of the portable amplifiers (which was partly unshielded); this eventually resulted in audio frequency cross-talk due to rectification in the amplifier vacuum tubes. The difficulty was remedied by inserting a trap circuit, consisting of a variable condenser in parallel with an inductance, in series with each side of the line, and tuning each trap to the interfering signal. The radio frequency current flowing in the line wires was thus greatly reduced and the cross-talk correspondingly lessened. Future models of portable amplifiers will be thoroly shielded to avoid this difficulty.

STUDIO OPERATION

In the organization of these stations the control operator is responsible for the general supervision and maintenance of the

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technical standard of the performance. He normally remains in the control room, but it is his duty to advise or correct the studio announcer in his placing of performers with respect to the microphone. For this purpose a "Dictograph" interphone system is provided, by means of which the control operator may signal or talk to the announcer; a lamp signal is used for attracting attention, or a buzzer signal may be used instead by throwing a switching key. The microphone used in this type of interphone is exceedingly sensitive, so that the announcer may talk to the control room in a very soft tone, during a performance, without disturbing the artist. The same interphone system connects to all other parts of the station.

Figure 29 shows some of the typical methods of placing performers with respect to the microphone. These are the result of experience in the WJZ studio and would probably not hold very strictly for other rooms. With all new performers, a trial placing is made according to experience, for the first number; this can be altered during the number, if necessary, by moving the microphone stand, or after the first number by suitable alteration of positions of performers. The announcer is usually familiar with the placing required for various types of performances and normally arranges the performers without any specific instructions from the control operator. The latter observes the effect of the announcer's placing, controls the degree of modulation, and makes an estimate, if possible, of the volume range the performers are capable of. Judging from this, he instructs the announcer to make such changes in placing as he deems necessary, so as to prevent blasting of the microphones or to secure a better "balance" (ratio of volume of soloist to accompanist or ratio of volume of various instruments in an ensemble).

It may appear that less work would have to be done with regard to proper placing of the performers if the microphones were not used relatively close up, being placed instead, for example, at the opposite end of the studio. In the latter case, the relative distance of the microphone from each of several performers would be nearly the same and there would apparently be less of a problem so far as proper "balance" is concerned. However, it has been found that this cannot be done for a number of reasons. First, the further away the microphone is from the performers, the greater is the proportion of sound which reaches it by reflection from the room walls, compared with that reaching it directly from the source of sound. These reflected sounds are

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generally distorted, since they not only are reflected in a variable fashion with respect to frequency, but interference phenomena occur between reflected sounds coming from various reflection points. Thus, it is found that the sounds as heard from a microphone located, say 20 feet (6 meters) from the source are more distorted than those heard when the microphone is placed relatively close.

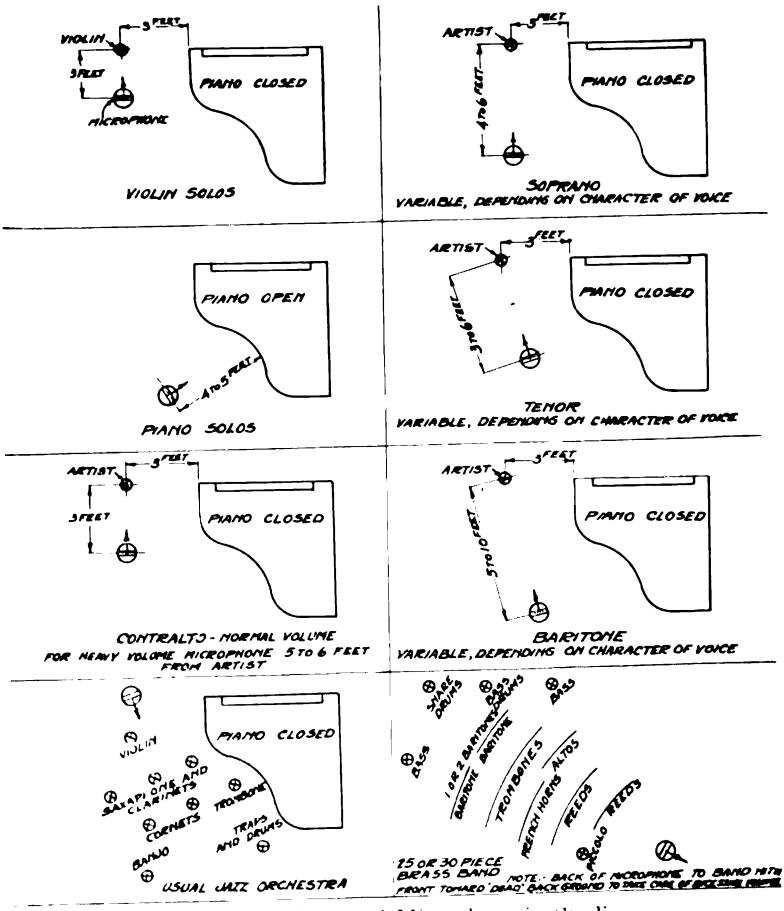

FIGURE 29 – Placing of Microphone in Studio

Second, the sounds reaching the microphone must be strong enough to give an output far exceeding the hiss due to the use of carbon, and this again necessitates fairly close placing with all performers except orchestras or large choruses.

Third, there is a certain advantage is giving the performers

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an object, such as the microphone stand, at which to direct their remarks, or music. A difficulty which all broadcasting stations experience is the lack of an audience to which the artists are accustomed. Usually this is helped by trying to have the radio listeners send in telegrams or make telephone calls during the performance, which are transmitted to the artist. We have tried to improve this situation further by placing a large mural painting on the wall of the studio, which is faced directly by the artist when broadcasting. In this, the painter has tried to depict some of the types of people who, the performer may imagine, are listening to him, and the vast spaces over which his voice or music may be traveling out to reach them. This is shown in the studio photograph, Figure 4.

One point that may be of interest is the variety of entertainment that is broadcast; Table I gives a list of typical program material transmitted from the studios at Broadcast Central during the first year of operation. (This table, as well as Table II, was very kindly supplied by Mr. Charles B. Popenoe, Program Director.)

TABLE I

STUDIO EVENTS
BROADCAST CENTRAL
May 15, 1923, to May 1, 1924

CLASSIFICATION	Number of Events
Alto Solos	3
Bands	48
Baritones	163
Bass Solos	32
Cellists	12
Concerts	157
Contraltos	104
Duets (Vocal and Instrumental)	12
Harpists	14
Miscellaneous Events: Celebrations, Music Lecture Recitals, Whistlers, Saxophones, Harmonicas, Languages, etc.	159

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(*Table I—Continued*)

Classification	Number of Events
Orchestras (including Dance Orchestras)	113
Piano	401
Poetry—Readings	100
Popular Songs	215
Quartettes	57
Songs (Sacred and Folk)	45
Sopranos	505
Stories (Bedtime)	294
String Music (Miscellaneous)	25
Talks	1336
Tenors	159
Trios	23
Violinists	186
Vocal Concerts	47
U. S. Army Bands	12
U. S. Navy Bands	12
U. S. Marine Band	1
Total	4235

Broadcasting From Points Outside the Studios

At Broadcast Central approximately 100 events per month are broadcast which emanate from points outside of the station, such as hotels, theatres, concert halls, and so on.

Table II shows the range and variety of subject matter that was handled from such points during the first year of operation of Broadcast Central. Special portable microphone and amplifier equipment has been developed for this work, in order to convert the sounds produced to electric currents of suitable character and intensity for transmission over a pair of wires to the broadcasting station, there to be radiated. A special wire line system has been erected in New York City for this service, a similar system is being installed in Washington, District of Columbia, and connections to station WGY of the General Electric Company, at Schenectady, New York (distant about 160 miles (256 km.)), as well as to station WRC at Washington, District of Columbia (distant about 220 miles (350 km.)) are also being provided; so that events at New York, Schenectady, Washington, or other cities enroute, may be broadcast simultaneously from these three stations if desired.

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TABLE II

Events Outside of Station Broadcast Central

May 15, 1923, to May 1, 1924

Classification	Total
Aeolian Hall Recitals	19
Banquets	70
Baseball Games	6
Boxing Bouts	8
Concerts (Vocal and Instrumental)	82
Church Services	100
Football Games	7
Band Concerts	16
Miscellaneous Events	69
N. Y. Board of Education Lectures	16
N. Y. University Lectures	44
Opera	3
Orchestra (Hotel and Dance)	348
Organ Recitals	116
Polo Games	3
Symphony Orchestra Concerts	23
Theatres (Plays and Musical Plays)	34
Town Hall Events (Addresses and Concerts)	14
Total	978

Wire Line System in New York City: The system in use has been installed by the Western Union Telegraph Company, in co-operation with the Radio Corporation of America, and consists partly of new facilities exclusively provided for this service and partly of existing facilities which are leased as required from time to time. The system is based to a considerable extent on the use of the elevated railways which traverse Manhattan Island (the principal borough of New York City) lengthwise, to carry trunk line cables. These cables have outlet terminal boxes and switchboards at important points, so that overhead twisted pair connections can be made from the trunk lines to points from which broadcasting is to be done. The system thus obtained is very flexible, makes it possible to lay trunk cables exclusively for this work at moderate expense, and also provides connections at a reasonable cost, since the work of running weatherproof twisted pair over the housetops, from the trunk cable terminal

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boxes to a desired point, can usually be done by one lineman in a few hours. Fortunately, also, the theatrical and hotel district, in which the great majority of events originate, is fairly restricted in area, and thus the trunk cables need not be very extensive. Connections to points in New York not reached by the regular broadcasting trunk cables are made via other facilities at the main terminal office of the Western Union Telegraph Company, situated at 24 Walker Street, by means of special pairs connecting to Broadcast Central. At this same office connections are also made to Schenectady, and to certain points outside of New York City.

The trunk lines so far installed on Manhattan Island are shown in Figure 30. These follow the elevated railways on Third, Sixth, and Ninth Avenues, with cross-connecting links. The trunk cables are terminated at small switchboards installed at the elevated railway stations, at certain points, so that connections of trunks may be made up as desired, and also so that a minimum length of "dead end" trunk cable is left connected in parallel to a working circuit. Terminal boxes at which overhead pairs may be bridged across any trunk line are installed about every 600 feet (200 meters) along the cables; in the less important districts they are placed only where it is likely that connections may be wanted.

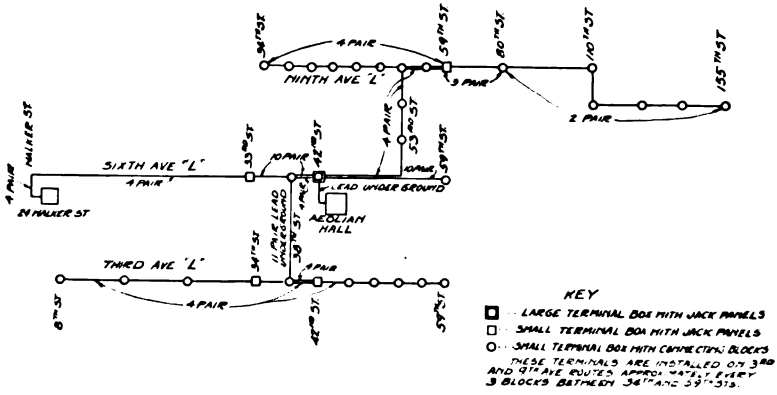

FIGURE 30—Wire Line System, Broadcast Central, New York City

An interesting point in connection with this elevated railway plant is its absolute freedom from induction, due to the motors of passing trains (the railways use 600 volts, direct current), tho the cable is only about five feet from the power rail. It was originally feared that lead-covered cable would be required, but

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the experience of the Western Union Company had been that lead-covered cable soon crystallized under the continuous vibration to which it is subjected on these railway trestles; hence a rubber-covered cable made up of rubber insulated pairs was recommended. However, the first installation of the rubber-covered cable soon proved that our fears of induction had been groundless, and it has accordingly been used thruout. The runs are sufficiently short so that the somewhat worse telephonic transmission qualities of rubber insulated wire as compared with the customary paper insulation do not matter to any appreciable extent.

Long Distance Wire Lines: Connections to points outside of New York City are obtained either via special facilities installed for the purpose by the Western Union Telegraph Company, or by the Postal Telegraph Company, or by lease of existing facilities. Figure 31 shows the long lines so far specially installed or in process of installation. Existing facilities available for broadcasting use are not shown. The Postal Telegraph Company is making connections between the New York and Washington stations, which enable events to be broadcast from these cities or also others en route, as indicated in the figure; while the Western Union Telegraph Company has furnished the connections to Schenectady.

The apparatus used in connection with simultaneous broadcasting thru the medium of this wire system will be described later on.

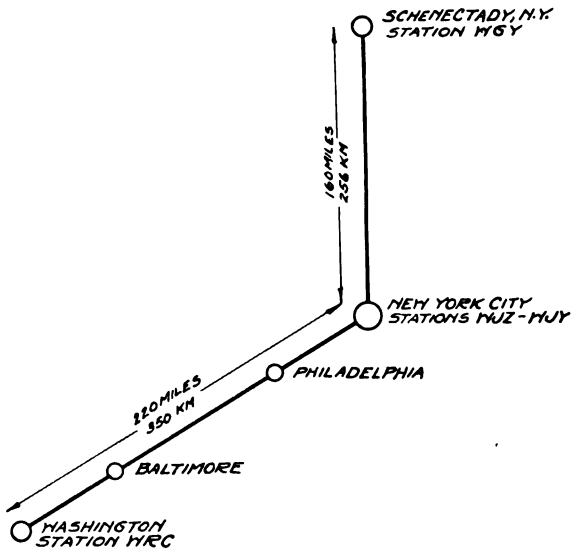

Figure 31—Long Distance Wire Lines for Broadcasting

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Portable Apparatus

In broadcasting from points outside of the station various types of portable amplifiers and battery boxes, microphones, microphone stands or housings, and an accessories kit for wire and tools are required.

Amplifiers: A standard set of equipment for the more usual types of work comprises the portable battery box and amplifier shown in the photograph, Figure 32, and its wiring diagram in Figure 33. The box at the left contains a 6-volt, 20-ampere

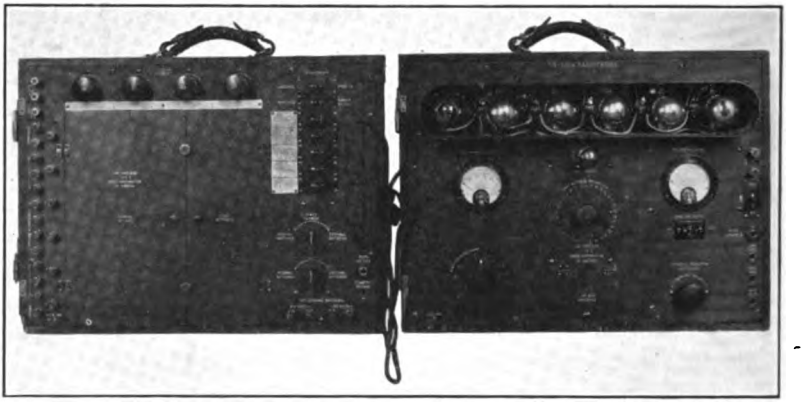

Figure 32—Standard Type of Portable Amplifier

hour battey for filament and microphone current, and five 20-volt "intermediate size" batteries forming a 100-volt battery for the plate voltage; here also (at the left) are mounted binding posts for connecting a total of five carbon double button microphones (called "concert" microphones), if desired, an announcing microphone, or the output terminals of a previous amplifier for example, for a condenser transmitter or line. At the right side are keys for switching from the announcing to the carbon concert microphones, or to the "condenser transmitter amplifier" or for connecting the carbon microphones in parallel. Cords and multiple contact plugs are provided for making connections from battery box to amplifier box. Switches are provided for connecting either the batteries contained within the box, or external ones in an emergency, to the amplifier. At the top of the battery box are four rheostat knobs which are used individually to regulate the microphone current of concert microphones numbered 1, 2, 3, and 4. These rheostats are a carbon compression type and permit the microphone current to be regulated smoothly

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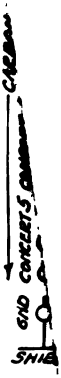

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and without noticeable noise, from a few milliamperes up to the rated current of 40 milliamperes. In this way, microphones placed, for example, on a stage and in the orchestra pit may be thrown in parallel, the volume contributed by each microphone (which is proportional to the microphone current) regulated individually, and a suitable balance of the voices from the stage with the orchestra accompaniment thus obtained. This feature is especially useful in broadcasting musical comedies or grand opera.

The right hand or amplifier box contains six tubes, along the top. Five of these are amplifier tubes, the sixth serves as a "volume indicator" (vacuum tube voltmeter indicating the voltage impressed on the telephone line). In the center of the panel is a "Gain Control," by means of which the amplification can be controlled by 20-percent steps from maximum to practically zero. This is merely a potentiometer between the plate circuit of the first tube and grid circuit of the second tube, as will be apparent from the wiring diagram. To either side of the gain control are meters for the microphone current and volume indication. Below the microphone current ammeter is a stepwise rheostat for roughly controlling microphone current, and below the gain control a compartment for housing a 9-volt battery for the volume indicator grid bias tapped at 4.5 volts for amplifier grid bias. At the right-hand end of the box are line binding posts, line switching key, and jacks for measuring various battery voltages, or currents, and for listening either directly across the line or in a special circuit fitted so as to permit the operator to regulate the intensity of the signal to suit himself, without affecting the outgoing signal intensity. A small automobile lamp, run off the 6-volt filament battery, and mounted above the gain control, has proved quite an essential feature, since it is frequently necessary to handle the amplifier in dark places (theatres).

Covers with padlocks are provided for both boxes, each box weighs about 35 pounds (16 kg.), and the two boxes with accessories kit can easily be transported by two men.

The maximum voltage amplification obtainable by this amplifier, from a 250-ohm input circuit (equivalent to a microphone) to a 250-ohm output circuit, is about 3,400 times (a gain of nearly 75 miles, in telephone terminology). However, we have never found it necessary to use this much, the usual practice being to work the amplifier at a step corresponding to a voltage amplification of about 200 times (49 miles). With this the announcer's speech (talking a distance of about 1 foot (30 cm.) from the mic-

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rophone) or other material being broadcast, with normal microphone placing (see Figure 41 showing typical placings for outside work) is brought to normal telephone level before being impressed on the line. (A line voltage of the order of 0.25 to 0.5 volt is usually employed.)

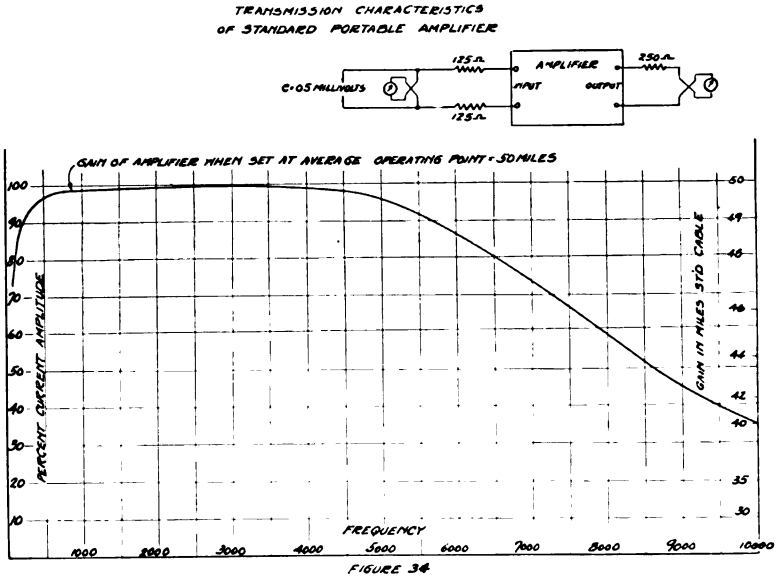

FIGURE 34—Transmission Characteristics of Standard Portable Amplifier

A frequency characteristic of one of these portable amplifiers is shown in Figure 34, with 250-ohm input and output circuits (resistance). The wide frequency range is obtained by the use of resistance amplification with the coupling elements calculated to transmit low frequencies with good efficiency, and carefully designed input and output transformers having high inductance windings with very low leakage.

A special type of portable amplifier for use in crowded places is shown in Figure 35. This is used at athletic contests, boxing matches, outdoor concerts or other events where only one or two seats can be had for the broadcasting staff. The operator can place this instrument between his knees (it is about 8 inches (20 cm.) square, at the base) and operate it readily, when seated in a chair among the audience, with the announcer in the seat next to him. If necessary, one man can both operate and announce. The amplifier apparatus is placed back of the sloping panel, and filament, bias, and plate batteries in the lower part of the box.

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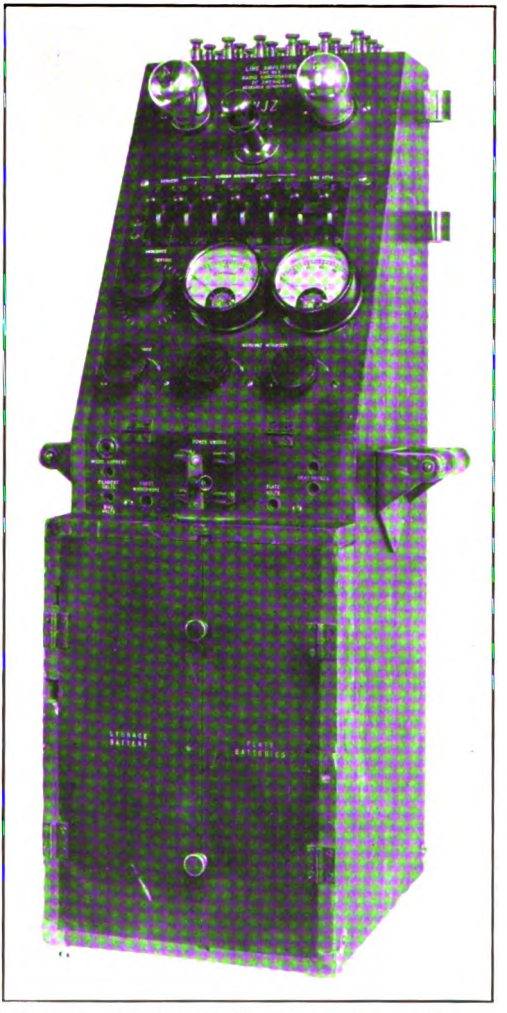

FIGURE 35—Special Type Portable Amplifier

The binding posts on the top are for announcing and other microphones and lines. There are posts for four "concert" or field microphones, one announcing microphone, and two telephone lines (one for the material to be broadcast and one for intercommunication). On the sloping panel are mounted, at the top, two vacuum tubes (the first one being a special high amplification constant tube, reactance-coupled to the second which is an output tube). Since the amplifier is used only for close talking or relatively loud sounds, less amplification is provided than in

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the standard amplifier; the total voltage amplification from a 250-ohm source to a 250-ohm output circuit is about 230 times (50 miles). Below the tubes are keys for the various microphones and a double key for interchanging broadcasting and intercommunication lines, so that if the broadcasting line should become noisy or fail the other line can be used instead. Below the keys are a gain control of the total amplification and ammeters showing the tube plate current and microphone current; the plate current ammeter is useful as an indicator of the presence of distortion, due to working of the output tube in excess of its linear amplication range, by the extent to which it flickers during speech.

The plate current is normally 7 milliamperes for both tubes and a flicker of a fraction of a milliampere is permissible. At the lower end of the sloping panel are three carbon compression rheostats for individually controlling the current of the announcing microphone and two of the field microphones. On the small vertical panel are a power switch, jacks for measuring the battery voltages, for connecting in an inter-communicating breast type telephone transmitter and for the operator's head telephones. One of the pair of telephones is connected across the inter-communicating line, the other to the broadcasting line, so that the operator can monitor on one line but still hear a call on the other.

In the practical operation of this equipment at an athletic contest, the field microphones are placed to pick up bands, cheering, announcements by field umpires, and so on. The individual volume controls and paralleling switches enable the operator to bring in these sounds as background to the announcing or after an announcement to gradually "fade in" or "fade out" the cheering or band music, thus giving a more artistic performance than merely the more or less dry announcing.

Another special amplifier is shown in Figures 36 and 37, and its wiring in Figure 38, for use with condenser transmitters.[10] Here it is necessary to shield the amplifier acoustically and electrically, because of the greater amplification necessary with condenser transmitters. The principal difficulties are acoustical, due to shocks or powerful low frequencies (such as the low tones of the organ), causing the tube elements to vibrate. Three tubes of the UV-201-A type are employed, the output of the amplifier being fed thru a step-down transformer to suitable input terminals on the standard portable amplifier previously described. Each

[10] References in connection with the condenser transmitter: E. C. Wente, "Physical Review," 2d series, volume 10, page 39, 1917; and volume 19, page 498, 1922.

I. B. Crandall, "Physical Review," 2d Series, volume 11, page 449, 1918.

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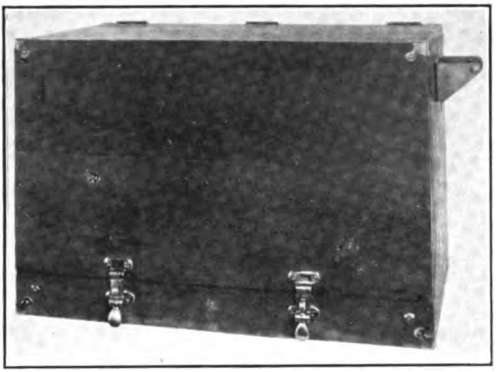

FIGURE 36—Condenser Transmitter Amplifier, Exterior

socket is mounted on soft rubber, and the three tubes enclosed in a metal case for electrical shielding. This case is hung on springs inside of a large wooden felt-lined box, for preventing the access of acoustic or mechanical vibrations. The condenser transmitter is used only where it is necessary to locate the pick-up device at a distance from the source, and where the sound is

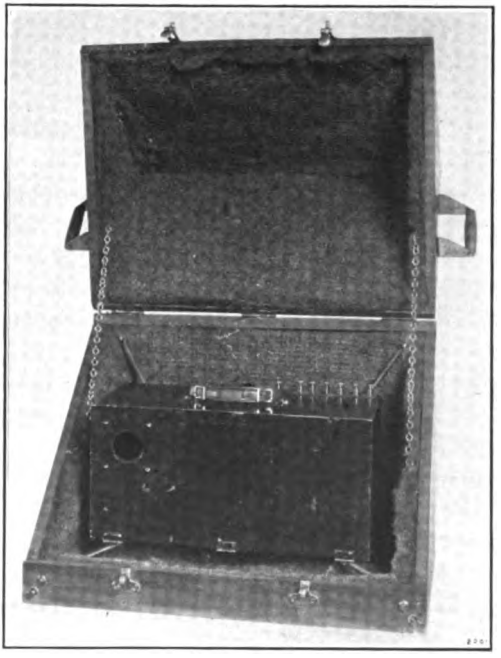

FIGURE 37—Condenser Transmitter Amplifier, Interior

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relatively weak, so that the hiss of the carbon transmitter would be noticeable to an undesirable extent. Such occasions arise at times in broadcasting large organs, or symphony orchestras, where it is desirable for good balance to have the microphone out some distance, because of the widespread area occupied by these sources. For general use, however, the condenser transmitter is undesirable because of the fact that long leads cannot be run to it from the amplifier, and the extra amplifier with batteries are an inconvenient extra burden.

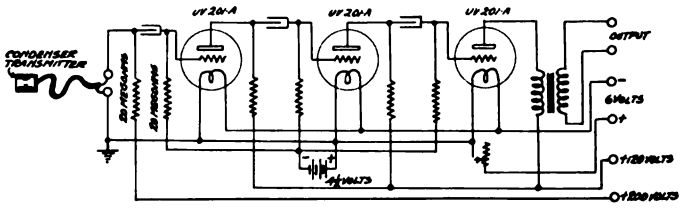

FIGURE 38—Condenser Transmitter Amplifier, Wiring Diagram

MICROPHONES AND MICROPHONE HOUSINGS: The microphones used on outside work are of the same double button carbon type as previously described (Figure 8), except that a modification is made in the diaphragm characteristics of these for certain kinds of work. In the microphones previously described, a diaphragm is used which is stretched so as to have a natural frequency of about 7,000 cycles, and heavily damped by means of an air damping chamber back of the diaphragm. Normally the spacing between the back of the damping chamber and diaphragm is about 0.001 inch (0.025 mm.). Under these conditions a certain ratio exists between the voltage generated by the microphone for a given sound pressure and the hiss due to the carbon, called the output-hiss ratio. For stage work it is sometimes desirable to use microphones having a greater output-hiss ratio than the normal, to pick up speech or music where the actors may be 15 to 20 feet (5 to 7 meters) from the microphone. This may be secured by making the microphones with a lower natural frequency (less tension on the diaphragm) and somewhat lower damping (greater air space). This sacrifices high frequencies somewhat, but not seriously. For other purposes a microphone going to the opposite extreme, namely, a lower output-hiss ratio is desirable. This is the case where the microphone is used on very loud sounds, which might cause the normal type to blast. Such a contingency occurs when the microphone is used in an-

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nouncing from boxing matches. Here the noise from the crowd is at times terrific, and the announcer must shout into the transmitter at the top of his voice to have his speech go over the noise. For this purpose the microphone can be made with heavier diaphragms, such as steel, so that the force impressed on the carbon is less for a given sound pressure than in the normal microphone.

These microphones are enclosed in various types of housings, except that where it is desired to render the microphone as inconspicuous as possible, it is hung without a housing. In the design of housings it is always necessary to be sure that the enclosure will not have a natural frequency in the audible range, otherwise distortion will result. It seems that most of the sounds we have to deal with, as they strike a microphone in its housing, contain a great many steep wave fronts, which shock-excite anything having a natural tendency to vibrate into oscillations at its natural frequency and damping. Hence resonance in a housing or in the microphone makes itself heard as a more or less continuous "ringing" along with the sounds desired. The best form of housing is one which has a minimum amount of solid wall, and in which the enclosing surfaces are perforated or covered with gauze as far as possible. In all cases the microphone is hung on springs or rubber bands.

A number of typical housings are shown in Figure 39. The smallest is used for general pick-up purposes, on the stage, for orchestras, public addresses, and so on. The next in size is for announcing, and is therefore made so as to be held in the hand. Then there is a support of adjustable height, for use with small orchestras, where it may be necessary to have the microphone from three to five feet (1 to 1.5 meters) above the floor. In Figure 40 a special housing of attractive appearance, designed[11] especially for use in broadcasting speeches from banquets, is shown (front and rear views). This housing, when standing on the table, brings the microphone to about the height of the speaker's chest; this gives a much better frequency response than having the microphone on the table, as it is found in the latter case that high frequencies are principally absent (perhaps due to greater absorption by the table coverings) and the speech sounds thick or muffled.

Operation Outside of Station

An event outside of the station is supervised from the con-

[11] The original model was designed by Mr. Edwin Field Sanford, Jr., a well-known sculptor of New York City.

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trol room in much the same way as a studio performance. The "outside crew," usually consisting of two men for a normal event, will report at the place from which the broadcasting is to be done, several hours in advance, set up their apparatus, run the microphone wiring, and test the line to the station which has been previously installed by a lineman. They can then test their choice of microphone placing, if suitable sounds are available (for example, in the case of a hotel orchestra), by listening to the output of the portable amplifier, and assure themselves that proper balance and correct output-hiss ratio are present. About

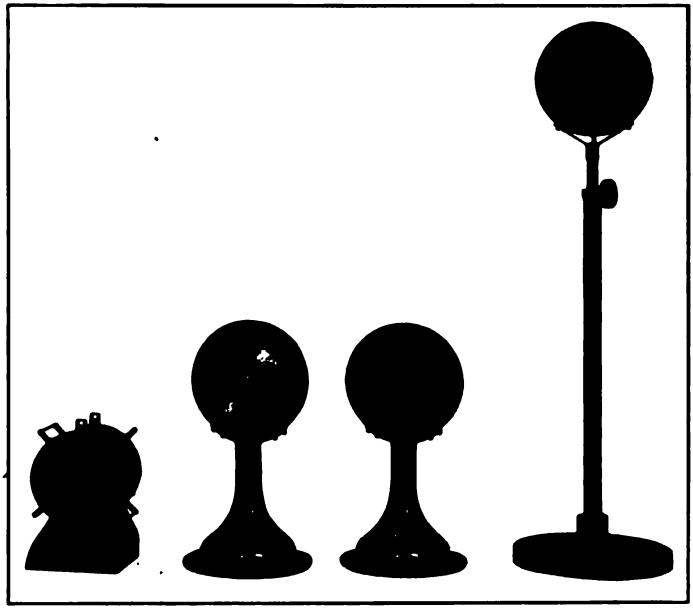

FIGURE 39—Typical Microphone Stands

five minutes before the event is due, they stand by with headphones across the broadcasting line and have their announcer and performers ready. The control operator then calls them and informs them approximately of the time he will be ready for them, the distant operators using the announcing microphone and amplifier to reply. At the moment the studio performance is completed the control man says "On the air" and switches the line to the input of the control room line amplifier. In the reverse case, where a transfer is to be made from the outside performance, a pair of telephone receivers is provided for the announcer in the studio by means of which he can listen to the material going out

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of the station; as soon as the distant announcer has announced the end of his program, the control operator switches the studio to the amplifier and at the same time the studio announcer takes up the program.

FIGURE 40—Special Microphone Stand, Designed by E. F. Sanford, Jr.

In the case of theatrical performances, one or more rehearsals are held before the broadcasting. This is done at actual performances, on previous matinees or evenings, with half a dozen microphones placed at strategic locations, on the stage and in the house, connected to the switches on a standard portable amplifier. The operator who is to handle the event finally chooses two microphone locations (usually one in or above the footlights and one in the orchestra), which will best pick up the performance, switching from one to the other if required, to take orchestra only, or sounds coming from the stage only. The switching is done by working the individual microphone volume

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controls, bringing up one while reducing the other, so that the listener is not aware of the change, as he would be from the click of a key.

Similar trial placings are made wherever the event is of a new character or likely to be difficult. However, the majority of the work cannot be carried out this way, as often the event only happens once (as in the case of athletic events) and set-ups must be made from previous experience with various contingencies anticipated if possible. This work outside the studio often calls for considerable resourcefulness and quick thinking on the part of the operating staff, if the performance is to go smoothly.

A number of microphone placings have become fairly well standardized, and serve as a guide when trials are not possible. Some of these are shown in Figure 41.

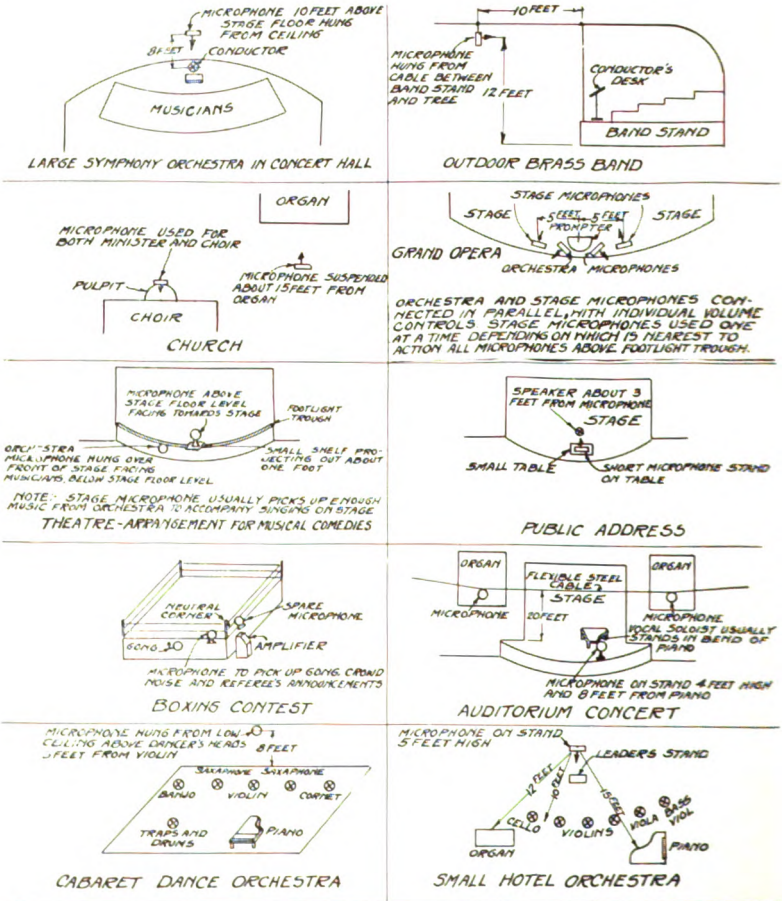

FIGURE 41—Microphone Placing in Outside-of-Studio Events

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In places outside of the station there is usually no means of controlling the reverberation time of the auditorium or the distribution of the sound. The general rule is to have the microphone fairly close up, and the distances indicated on the figure have been found to give quite good balance of orchestras, without the reverberation that would be heard at greater distances (for example, on the balcony rail in a theatre). It will be noted that for grand opera two microphones are used for the orchestra, one to take each of the two halves; while for the average orchestra a single microphone placed near the first violins usually picks up enough from the remaining instruments to give a sufficiently satisfactory balance. At any rate, until a pick-up device is invented that can be placed at a distance without showing excessive reverberation, relatively close placing with a multiplicity of microphones, paralleled when necessary, is likely to continue. With this close placing, apparently small differences in position will sometimes show surprisingly large differences in the character of the sound picked up. For example, a microphone placed in the footlights on a stage does not appear to get nearly as good a range of frequencies from speech on the stage as one placed *above* the footlight level and on a small shelf projecting out over the orchestra pit. Thus a certain amount of preliminary exploration is very desirable, particularly where abrupt changes in the contour of the house occur or where reflection or absorption of the sound waves is likely to take place.

One point in connection with work outside the station concerns the control of modulation. Where the sound volume to be transmitted is more or less uniform, as in the case of speech, the outside operator keeps his gain control set so that an approximately constant level of intensity is sent into the line, as indicated by the volume indicator, and the control operator at the station controls the percentage modulation as usual. However, with such work as grand opera, where the volume is constantly changing, the best method of control is to have the control room operator make an initial modulation setting corresponding to a certain level (on the announcer's speech, for example), and then have the operator at the theatre listen to the performance as it leaves the station's antenna, by means of a small portable receiving set, adjusting the gain on his portable amplifier so as to keep proper output or to avoid over-modulating the radio transmitter. Thus a guide to the average volume is obtained by the volume indicator, and to over-modulation by the sound of the performance as heard in the receiving set. After a very short time the operator

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soon learns to adjust the gain control so as to keep pace with the slow changes in volume, and to set his gain so as to avoid overmodulation on the occasional loud bursts. Of course, it is practically essential that an operator for this sort of work should have a good musical sense, and it is desirable that he be familiar with the operas to be broadcast.

Simultaneous Broadcasting and Special Wire Line Apparatus

In Figure 31 certain wire lines have been shown for linking the New York, Schenectady, and Washington stations for simultaneous broadcasting. To carry out this work necessitates the use of special apparatus, fitted to take energy from a line or from station equipment and transmit it out over a number of circuits simultaneously, without interaction between circuits, and without the possibility of an accident on one circuit affecting the others. At the same time it is desirable to arrange for the use of greater than ordinary telephone energy levels, if desired, so as to permit of a greater ratio of signal to disturbances (cross-talk or noise) on the line, and also to arrange for correction of line characteristics. All of this obviously involves the use of special vacuum tube amplifiers, and certain units of apparatus have been standardized on which can be arranged and connected (directly or via a switchboard) to meet various requirements. A number of such units are arranged on racks at each broadcasting station.

In Figure 42 (a) is shown a set-up for simultaneous broadcasting from a station or line to three outgoing lines. The apparatus is divided into units as shown, the input terminals of the preliminary amplifier and the output terminals of the multiple repeaters being connected to jacks on a telephone and telegraph switchboard, at which the various lines are terminated on jacks also, so that various connections can be made with plugs and cords.

When used from a station to lines, the preliminary amplifier is connected to the output from the main modulation control; hence, after a preliminary setting of volumes at the distant stations, the modulation at all stations is controlled simultaneously from the initiating station. A volume indicator is provided for use in making preliminary adjustments, but the normal functioning of the multiple amplifiers during operation is supervised by observing that the plate current ammeters of all the repeaters flicker in unison. A plate voltage of 500, and UV-202 tubes

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(normally rated at 5 watts as oscillators) are used, and with these a maximum of 5 volts (root-mean-square), audio frequency, may be impressed on the outgoing lines, should it be required.

With incoming signals, a repeating coil and line corrector (often termed "attenuation equalizer" in telephone practice) can be connected to any line, and the output of the corrector to the input of one of the line amplifiers in the control room amplifier rack, via the telephone switchboard.

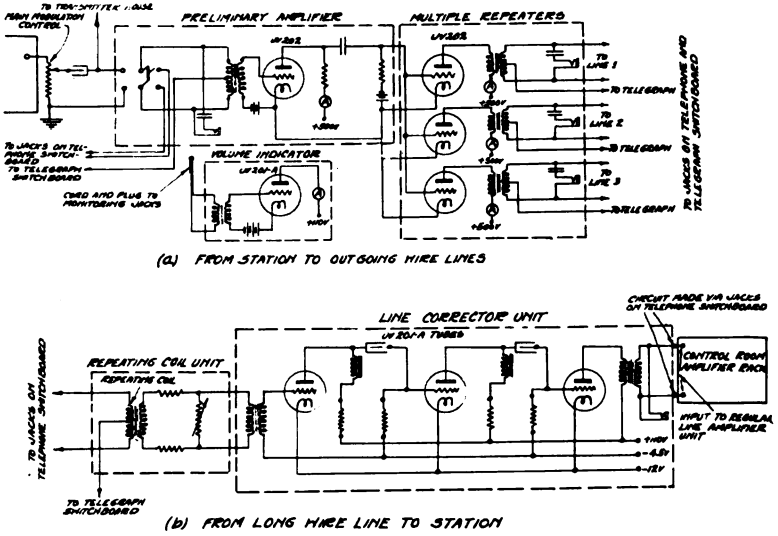

FIGURE 42—Certain Arrangements Used in Simultaneous Broadcasting

This is shown in Figure 42 (b). The function of a line corrector is to amplify various frequencies in inverse proportion to their loss on the line, so that the combined frequency transmission of line plus corrector is sensibly uniform over as wide a range as possible. In the particular corrector shown each tube and its associated circuit has a rising frequency amplification characteristic which when combined with the falling frequency transmission characteristic of the line gives uniform over-all transmission for a considerable frequency range. The required characteristic for any particular line is obtained by suitably proportioning the inductances and resistances in the plate circuits of the tubes. Similar arrangements can be made to correct transmission over various types of circuits, such as underground cable, open wire lines, and so on, and a corrector supplied to go with each circuit which is normally used. The three-tube corrector shown can also be used

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in sections, or other sections added, for greater or lesser lengths of the type of wire circuit which it is required to correct.

Communication between the various stations, while broadcasting is going on, is handled by means of telegraph over the same lines, the two wires of a pair being used in parallel against ground in accordance with standard practice.

Broadcasting Station at Washington, D. C.

This station, having the call letters WRC, is located at 14th Street and Park Road, N. W., the quarters being leased in an office building. The choice of location was necessitated by the fact that in the main hotel and theatre district of Washington it was not permitted to place antenna towers on the roofs of existing buildings, because of maximum building height restrictions. The site chosen is approximately 2 miles (3.2 km.) from this district, but connections to the principal possible sources of broadcasting are being made by means of underground cable installed by the Postal Telegraph Company.

The building in which the station is situated and the antenna towers are shown in Figure 43. The antenna towers are 100 feet (31 meters) high and the antenna is a "T" with flat top 160 feet (49 meters) long. Since this building is quite low, it was possible to place the studio, control and transmitter rooms on the same floor (about 20 feet (6.2 meters) below the roof), the antenna lead-in being run down the side of the building in the form of stiff copper tubing spaced away from the wall. The arrangement is less expensive in installation and operating cost than in the case of the New York station, where it was not possible to secure space on the upper floors of the building (close to the roof).

The arrangement of rooms (Figure 44) is similar to that in New York except that the transmitting sets and control apparatus are placed in the same room, so that one operator can handle all of this apparatus. Views of the operating and motor-generator rooms are shown in Figures 45, 46, and 47, and the studio in Figure 48. The equipment in general is the same as that in the New York station, and the acoustic treatment of the studio the same. A point that may be of some interest is the mounting of the motor-generator sets, to prevent vibration and noise from reaching the control room or other parts of the building; these are placed in a room having felt-lined walls and ceiling, each motor-generator set on a steel bed plate placed on a three-inch (7.6 cm.) cork mat; the cork is a few inches larger all around than

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the bed plate, and along each of its sides a piece of angle iron is placed, one side of the angle being secured to the floor and the other projecting several inches above the top of the bed plate.

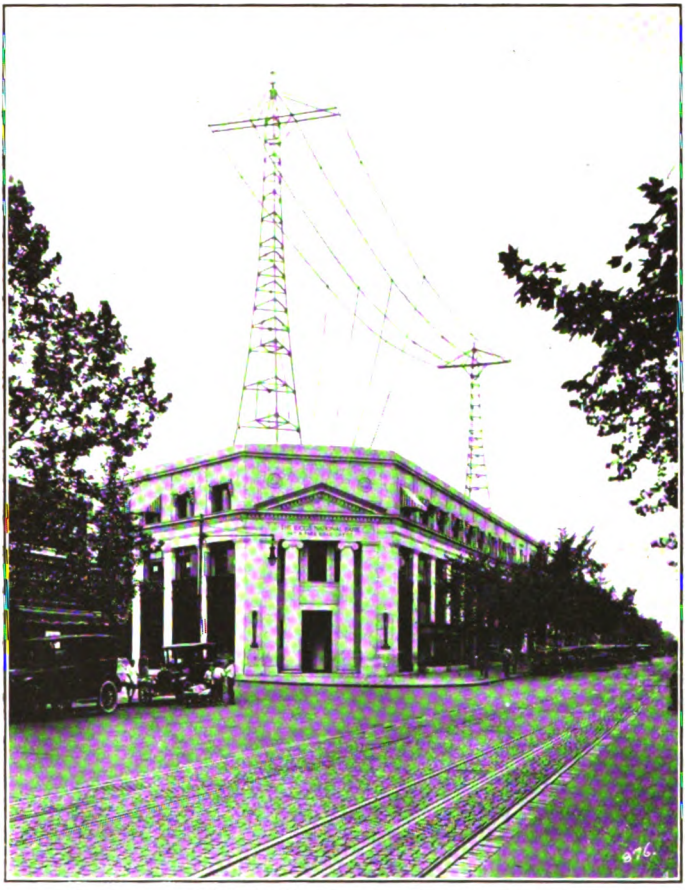

FIGURE 43—Site of Washington Broadcasting Station

Between this vertical side and the motor-generator base a cork strip is placed; thus the machine rests on cork, and is prevented from moving sidewise by the angle irons and cork strips. No holding down bolts are used on the machine base. Hence all vibration from the machine must travel thru cork, and the felt-lined room greatly reduces the noise generated within the room or transmitted thru the walls. The arrangement is quite successful, the noise of the machines being so low as to be negligible in the control room next door.

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The wire line system in Washington is not nearly so extensive as in New York, since Washington is a smaller city and there are a lesser number of points outside the studio from which events of broadcasting interest can be obtained. To reach these, a

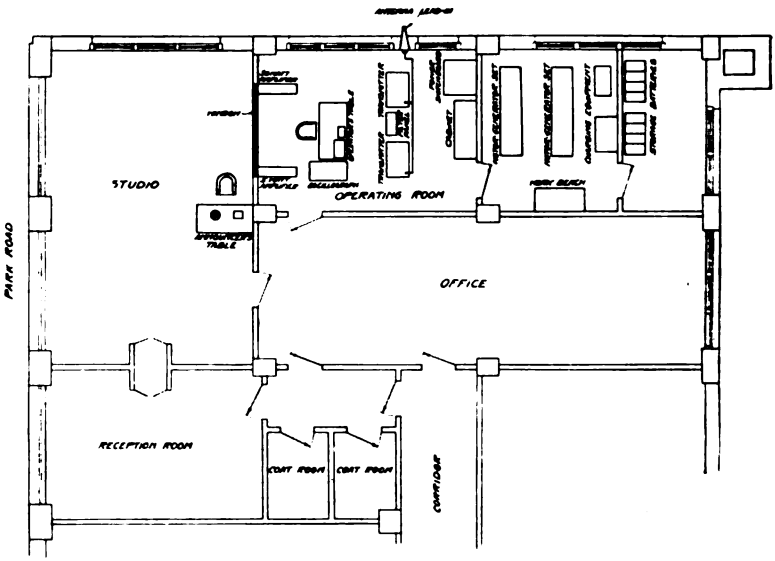

FIGURE 44—Plan of Rooms, Washington Station

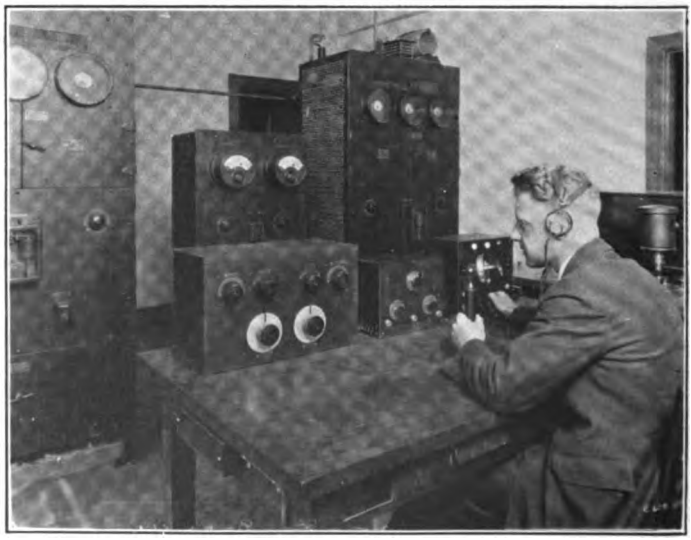

FIGURE 45—Washington Broadcasting Station Operating Room, Showing Control Desk, Duplicate Transmitters, and Filter Panel

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special 5-pair underground cable two miles (3.2 km.) long connects the station with the main terminal office of the Postal Telegraph Company, in the center of the hotel and theatre dis-

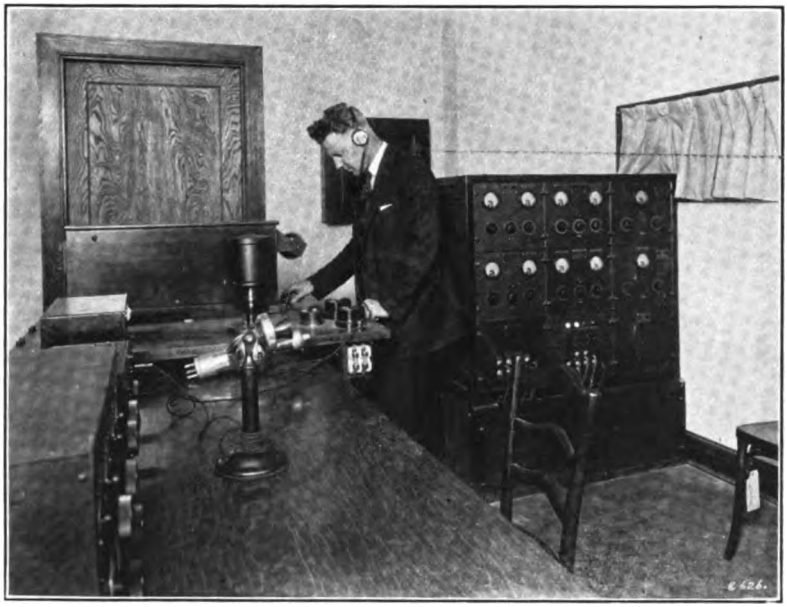

FIGURE 46—Washington Broadcasting Station Operating Room, Showing Control Room Amplifier Rack and Oscillograph

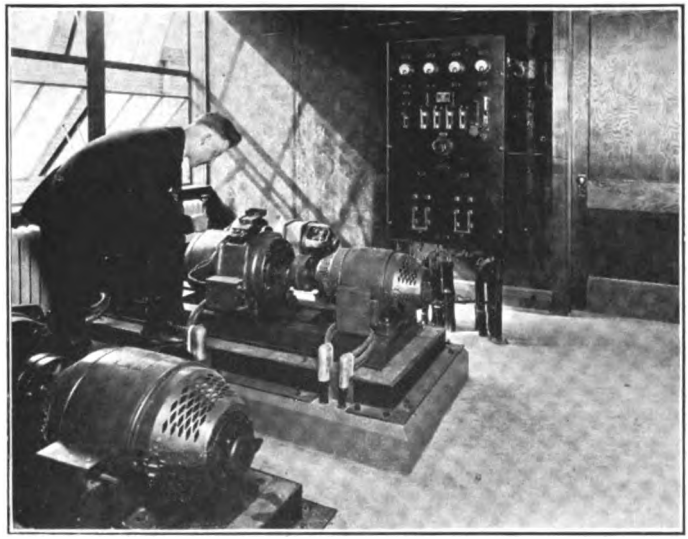

FIGURE 47—Washington Broadcasting Station, Motor Generator Room

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trict. Here connections are made as required, to various existing underground cables of this company in order to reach desired points. All of these cables consist of twisted pairs, and therefore are suitable for telephone use. One of the pairs in the cable to the station is permanently connected to the long distance line to New York, at the terminal office. The equipment used on the wires and the portable equipment is the same as previously described.

FIGURE 48—Washington Broadcasting Station, Studio

ACKNOWLEDGMENTS

In a project of this sort the responsibility for various portions of the work is naturally divided, and I wish gratefully to acknowledge the co-operation of the following gentlemen, all of the Radio Corporation of America, who were concerned in it:

Mr. J. H. Shannon, of the Design Division, Engineering Department, was in charge of the preparation of the detailed designs and specifications for the construction work.

Mr. S. C. Miller, of the Technical and Test Department, was responsible for the mechanical design of all of the portable equipment described in the paper, as well as certain station equipment. He also assisted in the experimental work and made all the drawings of this paper.

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Mr. Jesse Marsten, of the Technical and Test Department, was responsible for carrying out a good deal of the experimental work, alone or assisted by Mr. Miller.

Mr. Walter L. Tesch, now Engineer-in-Charge of the Washington Station, rendered valuable help in the initial testing and putting into service of the New York station.

Mr. Carl Dreher, now Engineer-in-Charge of the New York station, co-operated in all experimental work in a most friendly fashion, and furnished most of the data relative to operating experience given in the paper.

Mr. Raymond F. Guy, in charge of field (outside) operation at the New York station, has given valuable criticisms and suggestions concerning apparatus for this type of work, and also furnished data relative to experience gained therein.

Finally, the interest, encouragement, suggestions, and criticisms of Dr. Alfred N. Goldsmith, Chief Broadcast Engineer, have been of great value thruout the work.

SUMMARY: The general arrangements of the broadcast transmitting stations established by the Radio Corporation of America in New York and Washngton, D. C. (U. S. A.), are described. The station and portable apparatus, with wiring diagrams and frequency characteristics, are shown. Each station has an associated local wire line network, for broadcasting from points outside the studio, and long distance lines inter-connect the studios at New York, Washington, and Schenectady, N. Y., for simultaneous broadcasting. Equipment for this purpose is described. Various points in connection with practical operation in studio or outside broadcasting, such as supervision, microphone placing, and modulation control are dealt with. Experiences in two-channel transmission, at New York, are discussed.

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AN INTERNATIONAL COMPARISON OF RADIO WAVE-LENGTH STANDARDS BY MEANS OF PIEZO-ELECTRIC RESONATORS[1]

By

W. G. Cady

(Wesleyan University, Middletown, Connecticut)

While traveling last year in Italy, France, and England, the writer took advantage of the opportunity to make some wave-length comparisons between his piezo-electric resonators and the standard wave-meters or other high-frequency standards in those countries.[2] Six resonators were used of wave-lengths in the neighborhood of 200, 400, 1,300, 3,300, 10,300, and 20,800 meters. Four of these had been compared with the wave-meter at the Bureau of Standards in Washington in December, 1922, and all were sent to Washington for re-comparison at the close of the trip.

Observations were made in Italy at the Istituto Centrale di Radiotelegrafia ed Elletrotecnica del Genio Militare in Rome, and in the Istituto Elletrotecnico e Radiotelegrafico della Regia Marina at the Naval Academy in Livorno; in France, at the Etablissement Central du Materiel de la Radio-télégraphique Militaire in Paris; and in England at the National Physical Laboratory in Teddington, and the research laboratory of the Royal Aircraft Establishment at Farnborough.[3]

The dimensions of the resonators are as follows:[4]

G1, quartz rod $1.76 \times 1.13 \times 0.63$ mm.
G2, quartz rod $3.60 \times 1.40 \times 0.51$ mm.
G3, quartz rod $12.02 \times 2.16 \times 1.08$ mm.
G4, quartz rod $30.3 \times 2.64 \times 1.23$ mm.

[1] Communication from the International Union of Radiotelegraphy. Received by the Editor, June 3, 1924.

[2] For theory and description of the piezo-electric resonator, see Proceedings of The Institute of Radio Engineers, volume 10, page 83, April, 1922. A few corrections are given in the number for August, 1922, title page.

[3] I take this opportunity of expressing my thanks to Dr. G. Vanni, Director of the Institute in Rome, Lr. G. Vallauri, of the Italian Naval Academy, General Ferrié and Messrs. R. Jouaust and R. Mesny of the French Ministry of War, Mr. D. Dye of the National Physical Laboratory, and Major Le Fry of the Air Ministry in England, for the courtesies which they extended to me in their laboratories.

[4] For photographs of resonators, see Figures 7 and 9 in the paper in the Proceedings referred to above.

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These four rods were mounted together in a bakelite casing 8.9×1.7×1.6 cm.

F23, steel rod 90.5×9.75×3.16 mm., to the opposite sides of which were cemented quartz plates 8.9×8.95×1.13 mm.

F24, steel rod 180×9.23×3.14 mm., with quartz plates 13.4×9.0×1.18 mm.

Methods of Comparison

Four different methods of comparison were employed, the choice in each case depending upon the laboratory facilities available for the various wave-lengths.

Method A—The resonator is connected in parallel with the capacity of the wave-meter, the circuit of which must also contain an indicating instrument—for example, hot-wire meter or thermo-element of low resistance[5]. A tube generating circuit with fine frequency-regulation is loosely coupled to the wave-meter, and the two circuits are brought into resonance at approximately the natural frequency of the resonator. The generator frequency is then slowly varied one way or the other until a sudden and very sharply defined minimum in the wave-meter current indicates that the resonator frequency has been very exactly reached. Leaving the generator oscillating at this point, we remove the resonator and adjust the wave-meter to resonance with the generator as precisely as possible. In place of the wave-meter, an auxiliary tuned circuit containing resonator and indicating instrument may of course be used. The wave-meter itself is then compared by any desired method with the generator, after the latter has been adjusted to resonator frequency.

Method B—This is similar to Method A, save that the tuned secondary circuit, which may or may not be the wave-meter, contains a detector and telephone receiver in addition to the resonator. When the generator capacity is varied, resonance with the resonator is indicated by a click in the receiver, which usually comes at slightly different points on increasing and decreasing frequency. The mean of these is taken as the correct setting of the generator.

Method C—A telephone method, similar to Method B, but having in the generator circuit a small capacity and key in parallel with the tuning condenser. When the generator is adjusted to resonator frequency, the key is closed, causing the frequency to change abruptly by a small amount. A brief beat note is heard

[5] Proceedings of The Institute of Radio Engineers, previous citation, Figure 3.

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in the secondary circuit, caused by interference between this new frequency and the frequency due to the vibration of the resonator. When this note is of maximum loudness we know that the generator, before the key was pressed, was adjusted very precisely to resonator frequency.

METHOD D—This method makes use of the frequency-stabilizing property of the resonator.[6] The latter is connected in parallel with the tuning condenser of the generator, which must be of low power. The frequency is varied until the characteristic click in a receiver in the generator circuit, or in a tuned receiving circuit loosely coupled to it, is almost but not quite reached. If the secondary receiving circuit is regenerative, a beat note is heard, the pitch of which becomes practically constant when the stabilizing frequency of the resonator is reached. The generator frequency may then be compared with that of the wave-meter. For comparisons of highest precision, the difference between the stabilizing frequency and the natural vibration-frequency of the resonator must be taken into account. This difference is never over a tenth of a percent, and was disregarded in the present work.

Using Methods A or C, comparisons may be made which are accurate to within one part or less in 10,000. With Methods B and D, the order of accuracy is one part in 1,000. As to the constancy of the piezo-electric standards themselves, the following may be said: A variation in the distance between the brass electrodes, between which the crystal resonator lies, has a slight effect upon the natural frequency of vibration, as does also a shift in the position of the crystal when the air gap on each side is considerable. With the standards mentioned in this paper, such variations can hardly have amounted to more than one part in 1,000. Improvements in construction have been effected since these comparisons were made, which greatly reduce the possibility of chance variations in frequency.

Measurements of temperature coefficient indicate that an increase in temperature of 1° C. lowers the natural frequency of quartz resonators by one part in 200,000, and of steel resonators by one part in 10,000, approximately.

OBSERVATIONS

The comparisons made at the several laboratories will now be considered in detail.

[6] PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, previous citation, page 112.

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In Tables I to V, the symbol p stands for what we may term the "precision of comparison." It is an approximate estimate, expressed in percent, indicating the precision with which the wave-meter is probably tuned to the resonator. It is based on consideration of degree of coupling found necessary, accuracy with which the auxiliary generator set is tuned to the resonator, and precision of comparison between auxiliary generator set and wave-meter (or multi-vibrator). Except in Tables IV and V, p does *not* include the errors in wave-meter calibration. In most cases, the sharpness of tuning between generator and resonator was much greater than that between wave-meter and generator.

April 13-24, 1923. RADIOTELEGRAPHIC INSTITUTE, ROME.—Resonators G1, G2, G3, and G4 were compared with Wave-meter Number 6, Signals Experimental Establishment, Woolwich. A calibration of this instrument in terms of a Marconi standard, carried out some months later, indicated that all readings should be increased by two percent. This correction has been applied in the last column of Table I.

TABLE I

Resonator	Method	p	Wave-lengths in Meters		
			Observed	Mean	Corrected Mean
G1	B	0.5%	192	192	195.8
G2	B	0.5	388	388	395.7
G3	A	0.15	1250	1252	1277
G3	A		1255		
G3	B		1250		
G4	C	0.2	3190	3192	3256
G4	C		3195		

April 26-28, 1923. NAVAL ACADEMY, LIVORNO.—All resonators except G1 were compared with a standard Marconi wave-meter, Number 15,700. A later calibration of this wave-meter showed that all values of wave-length should be increased by 0.35 percent. The values so corrected are still probably somewhat too small, since in the resonator comparisons a thermo-element was in the wave-meter circuit. The calibration of the wave-meter consisted in a careful determination of its capacities and inductances, and took no account of the thermo-element.

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Unfortunately, a wave-meter which Dr. Vallauri had calibrated by means of a high-frequency alternator whose speed was controlled stroboscopically by a standard tuning fork, was not available for these comparisons. Observed and corrected values are recorded in Table II.

TABLE II

Resonator	Method	p	Wave-lengths in Meters	
			Observed	Corrected
G2	D	0.2%	385	386.4
G3	A	0.1	1222	1226
G3	D[8]	0.2	1240	1244
G4	C	0.15	3243	3254
F23	C	0.1	10278	10314
F24	C	0.1	20619	20691

June 15-18, 1923. MILITARY RADIOTELEGRAPHIC LABORATORY, PARIS.—Comparisons were made with Wave-meter Number 7, Modele Radiotélégraphique Militaire, System H. Armagnat, made by Précision Electrique, Paris. The circuits employed were those described by M. Mesny for the measurement of resistances at high frequency.[9] Extremely loose coupling was used thruout. Immediately upon the conclusion of the resonator comparisons, the wave-meter was compared with a standard tuning fork by Abraham's multi-vibrator method. The slight corrections obtained in this manner are included in the wave-lengtb values in Table III.

TABLE III

Resonator	Method	p	Wave-length, Meters	
G1	D	0.05%	197.2	
G2	A	0.02	394	Note on Table III: The two observations on G4 were separated by an interval of several days. In the absence of an explanation of the large discrepancy between them, the mean value, 3,264 m. has been chosen.
G2	D	0.06	394	
G3	A	0.05	1272	
G4	A	0.05	3260	
G4	A	0.1	3268	
F23	A	0.01	10,355	
F24	A	0.01	20,760	

[8] Thermo-element out, hence this value considered best.

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July 1-3, 1923. NATIONAL PHYSICAL LABORATORY, TEDDINGTON.—The writer was so fortunate as to be able to compare the resonators directly with the standard multi-vibrator, the fundamental frequency of which was determined by a tuning fork of 999.94 cycles per second. Method A was used thruout, an auxiliary generator set being carefully tuned to a thermo-electric circuit containing the resonator. The coupling was kept extremely loose—an important precaution when great accuracy is sought.

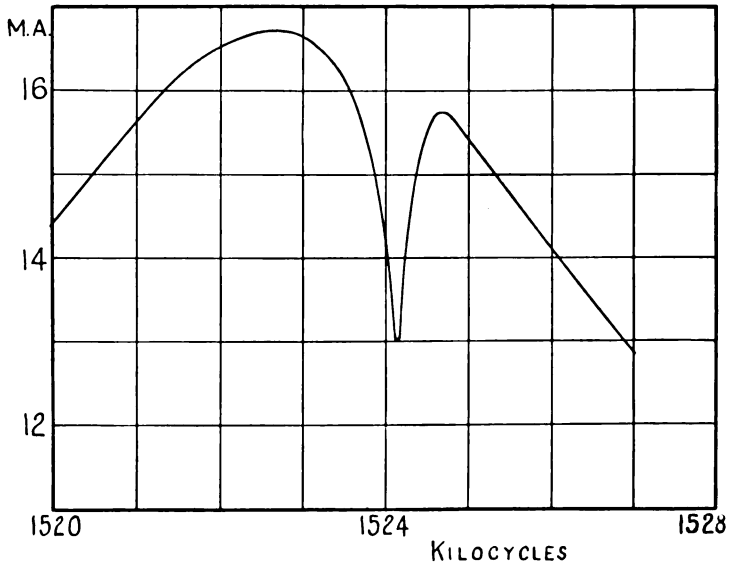

FIGURE 1—Resonance Curve for Quartz Resonator G1, which has a Frequency over 1,500 Kilocycles. The Resonance Point is at the Bottom of the "Crevasse"

The precision with which the generator could be adjusted to the resonator frequency is shown by the curve in Figure 1. Ordinates represent milliamperes in the thermo-electric circuit, and abscissas the impressed frequency. In the absence of the resonator, the curve would be an ordinary resonance curve, without the sharply-pointed "crevasse." When the resonator is in circuit, the curve suddenly develops a small and very sharp crevasse as the generator frequency is increased. The bottom of the crevasse is extremely sharp. It comes at the same setting of the generator condenser, whether the peak of the curve of electrical resonance occurs at this frequency or not, but

[9] Mesny, "L'Onde Electrique," volume 1, page 164, 1922.

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the effect is of course more pronounced the closer together the frequencies lie—that is, the better the electrical tuning of the thermo-electric circuit. In Figure 1 the electrical resonance comes at a slightly lower frequency than the mechanical resonance of the resonator. The figure also shows strikingly how much sharper the mechanical resonance was than the electrical resonance, in spite of the fact that the electric circuit was well designed and of low resistance. It should be noted that the fragment of quartz crystal producing this effect weighed only 3.3 milligrams.

Settings could be made to within 0.05 mmf. This corresponds to a precision of one part in 20,000, that is, of one centimeter in a 200-meter wave. Of course this has only to do with the precision in making settings. To measure the absolute frequency with this precision is another question. The multi-vibrator principle seems to be the only one at present making possible the measurement of absolute frequencies with this degree of accuracy.

A similar degree of precision (a few parts in 100,000), was attained with the other resonators. In each case, after the auxiliary generator circuit had been tuned to the frequency of the resonator, its frequency was determined by interpolation between selected harmonics of the standard multi-vibrator.

The fundamental frequency of the multi-vibrator was controlled, as has been stated, by a tuning fork of approximately 1,000 vibrations per second. The auxiliary generator set was brought alternately into synchronism with the quartz resonator, by the method outlined above, and with one of the multi-vibrator harmonics. In this way the effect of any drift in the frequency of the generator was eliminated.

The following information concerning the method of calibrating the generator set has been kindly furnished by Mr. Dye, of the National Physical Laboratory:

"The calibration of the generator set is carried out by adjustment of an open scale standard air condenser forming a small part of the total capacity of the generator oscillatory circuit.

"When the selector condenser of the multi-vibrator selector is set so as to select the desired harmonic having a frequency near that under measurement, a beat tone is heard in the amplifier receiving the two frequencies. By adjustment of the generator frequency this beat tone can be reduced to zero frequency, when the generator frequency becomes synchronous with that of the selected harmonic.

"Fiducial points are thus obtained at frequencies of, for ex-

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ample, 30, 31, 32 . . . kilocycles per second. It will be seen that in this region of frequency these points differ in frequency by three percent, which, tho not large, does throw considerable resposibility on the accuracy and uniformity of the scale of the standard condenser when an accuracy of one or two parts in ten thousand is required. It is fortunate, however, that a number of subsidiary synchronizations can be obtained at frequencies intermediate between the main harmonics. For example, when the generator is set at frequencies close to 30.250, 30.333, 30.500, 30.667 and 30.800 kilocycles per second, slow beats may also be heard and can be reduced to zero rate by very precise adjustment of the source.

"The explanation of these beats is, of course, that in the multivibrator selector circuit currents at the frequencies of the neighboring harmonics 29 and 31 are present. When the generator is set at, say, 30.333, the difference tone 333.3 cycles per second between its frequency and the 30th harmonic of the multivibrator is heard, and also the difference tone 666.7 cycles per second between the generator and the 31st harmonic. The second harmonic of the difference tone 333.3 cycles per second then interferes with the 666.7 cycles per second difference tone and gives the slow beat by which the final setting is actually made.

"The setting on the generator condenser is extremely precise for these sub-harmonics and the resulting accuracy is therefore very great."

For example, consider the case where one is obtaining a beat between the 4th harmonic of the 200 beat tone corresponding to one main multi-vibrator harmonic, and the 800 beat tone corresponding to the next multi-vibrator harmonic, that is, a generator frequency of 30,200. If now the generator frequency is altered to 30,201, the 200 beat tone becomes 201 and the 4th harmonic of 201 is 804; also the 800 beat tone becomes 799, the beats between these are 5 per second for a change of one cycle per second in the source. These subsidiary harmonic beats are very valuable for interpolating purposes, and remove nearly all the responsibility from the frequency-adjusting condenser in the generator circuit."

For the resonators of highest frequency, a harmonic of a harmonic was employed in calibrating the generator set. That is, by the use of the selector circuit of the multi-vibrator, one particular harmonic of the 1,000-cycle fundamental was amplified, and a group of its harmonics used for the calibration. Thus in

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the case of the resonator G1, of over 1,500-kilocycles frequency, it was necessary to extend the range to the 76th harmonic of a fundamental of 20,000 per second, this latter being in turn the 20th harmonic of the tuning-fork frequency. Synchronism was obtained between the generator and sub-harmonics of frequencies 1,523, 1,524, 1,524.5, 1,525, and 1,526 kilocycles, as well as the main harmonics (76th and 77th), 1,520 and 1,540 kilocycles, respectively.

The results obtained at the National Physical Laboratory are given in Table VI. The precision of comparison in each case is estimated to be a few parts in 100,000.

July 20, 1923. Royal Aircraft Establishment, Farnborough.—Resonator G4 was compared with the standard wave-meter by methods A and D, and a mean value of 3,275 meters found for the wave-length, the order of precision being 0.1 percent.

At the Bureau of Standards in Washington the frequencies of Resonators F23, F24, and G3 were measured in December, 1922, and in September, 1923, the frequencies of all six were measured. Method A was used, and the auxiliary generator set was compared with the standard wave-meter. The precision was estimated at the Bureau of Standards to lie between 0.1 and 0.2 percent, probably nearer to 0.1 percent, and this figure includes the precision of the wave-meter itself. As the writer did not make the comparisons in person, he is unable to estimate the probable precision of the piezo-electric comparisons by themselves, p, as distinguished from the probable accuracy as the wavemeter, hence the value of p in Table IV includes the estimated wave-meter accuracy.

TABLE IV

Date	Resonator	Wave-length in Meters	p
December, 1922	G3	1,271	0.1%
December, 1922	F23	10,390	0.1
December, 1922	F24	20,800	0.1
September, 1923	G1	196.5	0.1
September, 1923	G2	393.1	0.1
September, 1923	G3	1,270	0.1
September, 1923	G4	3,277	0.1
September, 1923	F23	10,400	0.1
September, 1923	F24	20,790	0.1

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In Table VI, the mean values from Table IV are reproduced, converted into kilocycles.

The Bureau of Standards wavemeter is calibrated for the lower frequencies in terms of a standard tuning fork, for the higher frequencies by means of stationary waves on parallel wires. Over the range where the frequencies overlap, the values check to within 0.1 percent.

In January, 1923, Professor G. W. Pierce kindly measured the frequencies of resonators G1, G2, G3, and G4, by comparison with the multi-vibrator in the CRUFT LABORATORY at Harvard University. In October, 1923, these observations were repeated, and data were also obtained for F23. Method D was employed. Following are the values obtained, a slight correction for frequency of standard tuning fork having been applied:

TABLE V

Date	Resonator	Wave-length in Meters	p
January, 1923	G1	196.7	0.1%
January, 1923	G2	393.6	0.1
January, 1923	G3	1,273	0.1
January, 1923	G4	3,272	0.1
October, 1923	G1	196.5	0.1
October, 1923	G2	392.9	0.1
October, 1923	G3	1,271	0.1
October, 1923	G4	3,266	0.1
October, 1923	F23	10,363	0.1

Mean values from this table are given in the general summary in Table VI, converted into kilocycles. The same remarks concerning the precision of comparison p apply to Table V as to Table IV.

SUMMARY AND CONCLUSION

The entire series of comparisons is now summarized in Table VI, wave-lengths being converted into kilocycles. The velocity of electromagnetic waves is assumed to be 3×10^{10} centimeters per second.

In computing the "weighted means," only those values based ultimately upon the frequency of a standard tuning fork (or, at

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the Bureau of Standards, upon waves on parallel wires) are considered. Inasmuch as the comparisons at Teddington were made directly with the multi-vibrator, weight 3 is attached to these values. In Paris, one day elapsed between the comparison of the wave-meter with the resonators and that with the multi-vibrator, hence weight 2 is assigned in this case. Values obtained at the Bureau of Standards and the Cruft Laboratory receive each weight 1.

TABLE VI

Station	G1	G2	G3	G4	F23	F24	Weight
Rome	1,532	758.2	234.9	92.14	.....	.....	.....
Livorno	.....	776.4	241.2	92.19	29.09	14.49	.....
Paris	1,521	761.4	235.8	91.91	28.972	14.451	2
Teddington	1,524.1	761.82	235.90	91.480	28.960	14.421	3
Farnborough	.....	.....	.....	91.60	.....	.....	.....
Bureau Stand	1,526	762.7	236.1	91.55	28.835	14.420	1
Cruft Lab	1,526	762.9	235.9	91.78	28.948	.....	1
Weighted means	1,523.8	761.99	235.90	91.656	28.944	14.431	.....
Probable error of weighted means in percent	0.032	0.018	0.009	0.051	0.040	0.027	

The weighted means are shown at the bottom of Table VI, as well as the probable error of each, computed in the usual way from the values used in deriving the weighted means. This "probable error" should be regarded only as a means for judging the closeness of agreement between the observed values at the different stations. Owing to the fact that in some cases different methods of comparison were employed at different stations, and also owing to possible changes in the natural frequencies of the resonators resulting from transportation and so on, the values should not be taken too confidently as an exact measure of the agreement between the wave-length standards in the countries visited. Nevertheless, the writer ventures the belief that they may not be entirely valueless in this respect.

The deviations of the values obtained at each station from the weighted mean are given in Table VII.

Considering only Paris, Teddington, Bureau of Standards, and Cruft Laboratory, we find the greatest single deviation to be 0.38 percent, while the average of all deviations in Table VII for these four stations is 0.11 percent. This is of the same order of magnitude as the estimated "precision of comparison," and it indicates that the wave-length standards in the four laboratories

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named are in agreement, on the average, to within one part in a thousand.

TABLE VII

DEVIATION, IN PERCENT, FROM WEIGHTED MEANS

Station	G1	G2	G3	G4	F23	F24
Rome	+0.54	−0.50	−0.43	+0.52	.....	.....
Livorno	.....	+1.5	+2.2	+0.58	+0.50	+0.41
Paris	−0.18	−0.08	−0.038	+0.27	+0.097	+0.14
Teddington	+0.02	−0.022	+0.000	−0.192	+0.055	−0.07
Farnborough	.....	.....	.....	−0.061	.....	.....
Bureau Stand	+0.14	+0.09	+0.08	−0.12	−0.38	−0.076
Cruft Lab	+0.14	+0.12	0.00	+0.14	−0.01	.....

If the available time and facilities had been such that all comparisons could have been made by the most precise method (Method A), more importance could be attached to the values of these deviations.

In constancy and sharpness of tuning, as well as in simplicity, the piezo-electric standards are already superior to any type of wave-meter in use at present, except the multi-vibrator. Further refinements in mounting and in methods of comparison will be described in a later paper. In cases where a precision of one or two parts in a thousand is sufficient, comparisons with the resonators can be made very easily and quickly, without other auxiliary apparatus than a detector and telephone receiver.

Wesleyan University, Middletown, Connecticut,
June 2, 1924.

SUMMARY: Piezo-electric wave length standards of the quartz or quartz-steel types previously described by the author have been compared with standard wave meters or multi-vibrators in the United States, Italy, France, and England. Six standards were used, of frequencies ranging from 14 to 1,500 kilocycles. The methods of comparison are described, and data obtained at the various laboratories presented. The tabulated results show that the average deviation from the weighted means, for all comparisons, is about 0.1 percent. It is shown that by the simplest methods of comparison a precision of 0.1 percent is attained, while in a comparison with the multi-vibrator, by more elaborate means, the precision is a few parts in 100,000.

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CORRECTION FACTOR FOR THE PARALLEL WIRE SYSTEM USED IN ABSOLUTE RADIO FREQUENCY STANDARDIZATION*

By

August Hund

(Electrical Engineer, Bureau of Standards, Washington, D. C.)

A pair of parallel wires was used in the early days of radio by Hertz, Lecher, Lodge, and others for demonstrating stationary electromagnetic waves. With the development of radio-frequency generating sets producing sinusoidal currents, this experiment became simpler, so that voltage nodes and the corresponding current antinodes can now be located with ordinary thermoelectric instruments with a high degree of precision.

It is the purpose of this paper to give the formulas and the corrections needed when the parallel wire system is used for primary frequency standardization.[1] In the parallel wire system[2] used by the Bureau of Standards the basis of frequency standardization involves the direct measurement of the wave lengths of very short[3] waves (4 to 20 meters in length) on a pair of parallel wires as indicated in Figure 1. The wave length λ is found by moving a shunted thermo-galvanometer G, which is suspended between the two wires, along the wires until it shows a maximum current. This galvanometer arrangement has a very low effective resistance. The point of maximum current is marked on the wires as position I-I and the galvanometer moved still further along the wires until a second current maximum II-II is indicated. The distance I-II between these points of successive current maxima is one-half wave length. If the paral-

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[1] The complete theory is given by the author in a Bureau of Standards Scientific Paper now in press.

[2] A detailed description of the arrangement and the apparatus was given by F. W. Dunmore and F. H. Engel in the Proceedings of The Institute of Radio Engineers, page 467, October, 1923.

[3] For standardizing lower frequencies, a local generating set producing currents of lower frequencies is loosely coupled to the detector and the settings of the frequency meter are compared to the frequency of the short waves by means of the harmonics of the low frequency source.

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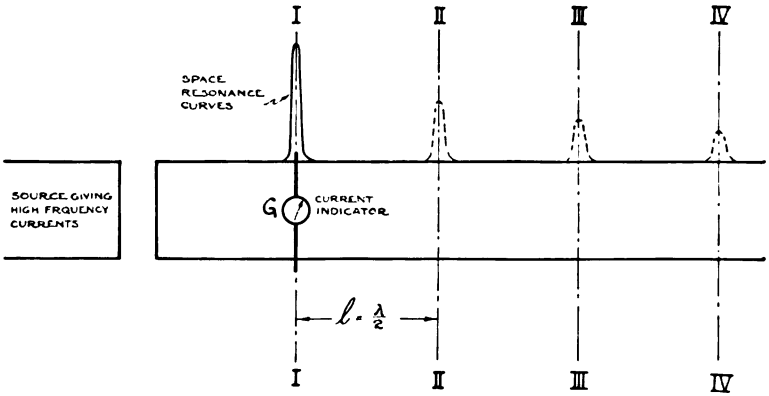

FIGURE 1—The Parallel Wire System of the Bureau of Standards Indicating the Space Resonance Curves as Determined with the Ammeter Bridge

lel wires are sufficiently long a number of such points, as III-III, IV-IV, and so on, may be found. The actual determinations are made with the positions I-I and II-II only since the distance $l = \frac{\lambda}{2}$ can be measured most accurately and the value obtained can be shown to be most reliable. Using a small condenser across the input terminals shifts the two maximum settings closer to the input end without changing the distance between them. The frequency f may be calculated from the formula.

$$f = \frac{v}{2l}$$

where v is the velocity of propagation along the parallel wires. The theory shows that the wave length λ set up on the parallel wire system is just $2l$, that is, twice the distance between two consecutive maximum settings except for a second order small quantity, but the velocity v with which the wave is propagated in the direction of the parallel wires is not the velocity of light ($v_o = 2.9982 \times 10^{10}$ cm./sec.) but a smaller velocity

$$v = v_o(1 - \Delta)$$

This velocity divided by the true wave length λ must be equal to the frequency f.

The frequency f in kilocycles per second is therefore calculated from the formula

$$f = \frac{v_o}{2l}(1 - \Delta)$$

where v_o is the velocity of light. Therefore,

$$f = \frac{1.4991 \times 10^5}{l}(1 - \Delta) \qquad (1)$$

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where the distance l between two consecutive maximum settings is expressed in meters. Figure 2 gives a plot for the small quantity Δ for different values of l. This curve is calculated for a parallel wire system using copper wire, 0.145 cm. diameter (number 15 American Wire Gauge) with a spacing of 4.2 cm. between centers.

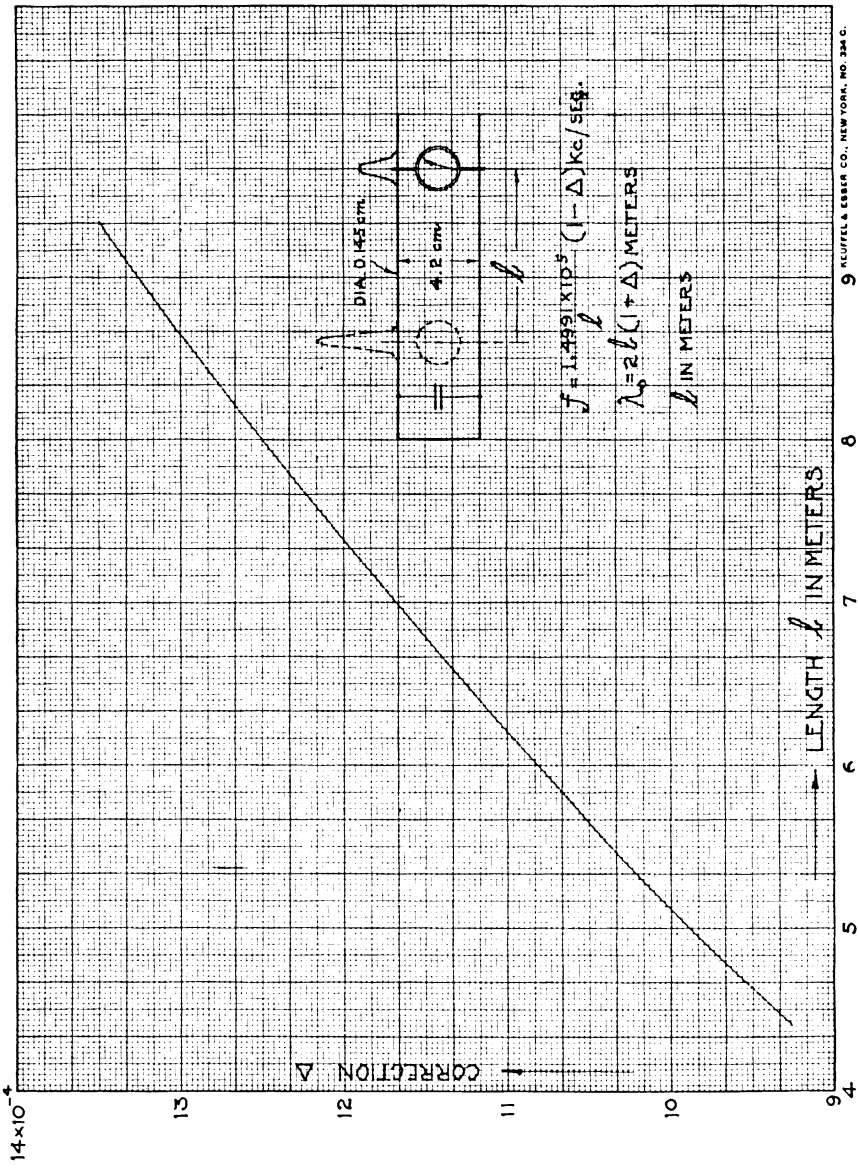

FIGURE 2.—Curve Giving the Correction Factor Δ for Different Values of l and Fixed Dimensions of the Parallel Wire System

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The calculation is based on the expression [4].

$$\Delta = \frac{\sqrt{r_o}}{8 l n B \sqrt{\omega\left[1-\left(\frac{d}{a}\right)^2\right]}} \qquad (2)$$

where

$$B = \frac{1+\sqrt{1-\left(\frac{d}{a}\right)^2}}{\frac{d}{a}}$$

which for normal spacings between the centers of the wires which are as a rule large compared with the diameter d of the wire gives practically $B = \frac{2a}{d}$

a = distance between the centers of the wires.

d = diameter of the wire.

r_o = the direct current resistance per centimeter length of the double line expressed in electromagnetic c. g. s. units (10^9 e. m. u. = 1 ohm).

$\omega = 2\pi f$.

f = frequency in cycles per second.

Formula (2) is applied as follows: Using for example the parallel wire system of the Bureau of Standards for a frequency $f = 2 \times 10^7$ cycles per second, a diameter $d = 0.145$ cm., and a spacing $a = 4.2$ cm., $\frac{d}{a} = 0.0345$; $\left(\frac{d}{a}\right)^2 = 0.00119$ which is in this particular case negligible compared with unity. Hence

$$B = \frac{2a}{d} = 57.92.$$

Assuming the resistivity of copper equal to 1,600 c. g. s. units, we find for one centimeter length of the parallel wire system

$$\sqrt{r_o} = \sqrt{\frac{2 \times 1600}{\frac{\pi}{2} \times 0.145^2}} = 440 \text{ electromagnetic c.g.s. units}$$

and $\Delta = 0.00121$, which shows that the frequency would be about 1/10 of a percent too high if the correction term Δ were neglected in equation (1). The result has been substantiated by experiment; frequencies calculated with formula (1) agreed with results

[4] This formula utilizes among other quantities an unpublished high frequency formula for the inductance of the parallel wire system which is due to Dr. Chester Snow of the Bureau of Standards.

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obtained by an entirely different method of frequency standardization using Lissajous figures and a standard tuning fork.[5]

Department of Commerce,
Washington, D. C.

SUMMARY: The above paper describes briefly calculations required in the parallel wire system used by the Bureau of Standards for radio frequency standardization. Formulas involving a correction term are given by means of which the standard frequency can be calculated from the distance between two consecutive maximum settings along the parallel wire system. A numerical example is added, showing the magnitude of the correction term.

[5] This method is described in a Bureau of Standards Scientific Paper now in press.

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ON THE RADIATION RESISTANCE OF A SIMPLE VERTICAL ANTENNA AT WAVE LENGTHS BELOW THE FUNDAMENTAL*

By

Stuart Ballantine

(John Tyndall Scholar in Physics, Harvard University, Cambridge Massachusetts)

The calculation of the radiation resistance of a simple vertical antenna erected over a perfectly conducting plane earth has been discussed by several investigators, including Abraham[1] and Pierce.[2] Abraham regards the wire as equivalent to an ellipsoid of revolution the major axis of which is very much longer than its minor axis, and rigorously investigates the oscillation of charge by applying Maxwell's equations to it in suitable curvilinear coordinates; he thus derives values for the damping due to radiation at certain definite wave lengths corresponding to the fundamental and the several harmonics. Pierce has extended the calculation to wave lengths above the fundamental by making the assumption that the variation of the current amplitude along the antenna remains sinusoidal when an inductance is included at the base; he is then able to calculate the radiated power by integration. The value of the resistance at the fundamental (36.57 ohms) obtained by Pierce is in agreement with that rigorously derived by Abraham, but it should be noted that this does not imply that the assumed current distribution is correct for wave lengths other than the fundamental; there are, in fact, good reasons to suspect that the current distribution is not of such simple sinusoidal type, but, because of the non-uniformity of the inductance and capacity per unit length, would be given by more complicated functions, such as Bessel's I_m and K_m functions. Stone and Roos discussed this twenty years ago and the subject has lately been revived by Press in these Proceedings and in the London "Electrician." I do not propose to examine this point

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[1] M. Abraham: "Ann. der Phys.," 66, page 435 (1898); "Math. Ann.," 55, page 81, (1899); "Phys. Zeit.," 2, page 329, (1901).

[2] G. W. Pierce: "Proc. Amer. Acad.," 52, page 192, (1916), or "Electric Oscillation and Electric Waves," page 435 (New York, 1920).

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critically in the present short paper, but merely to extend the calculations of radiation resistance to wave lengths below the fundamental, making use of Pierce's assumption of sinusoidal current distribution. In view of the uncertainty of this assumption the extension may appear to be of little final value; nevertheless the results are by no means trivial, since they permit the examination of certain important questions, which, if we wait for a really rigorous theory, will probably remain unsolved. One such question, which I propose to discuss in a separate paper, is that relating to the best transmitting wave length for a vertical antenna. It turns out that this wave length, for highly conducting earth, lies below the fundamental, hence the importance of these calculations. Aside from this it is well to have them on permanent record in the hope that experimental data may be available for comparison later.

Method of Calculation

The method here used is the classical one derived from the electron theory, which furnishes an easy means of finding the electric and magnetic forces by integration when the distribution and motion of the charges is known. Pierce's doublet method is equivalent to this, but somewhat less elegant mathematically. The intregation extends over the known distribution of current in the antenna wire and the current which spreads out radially over the surface of the earth. Since the earth is assumed to be perfectly conducting, this surface current may be represented by an image of the antenna extending below the earth. The integration is then simplified and extends over the antenna and its image.

For convenience the set of spherical coordinates shown in Figure 1 will be employed. We shall also use Gaussian units,

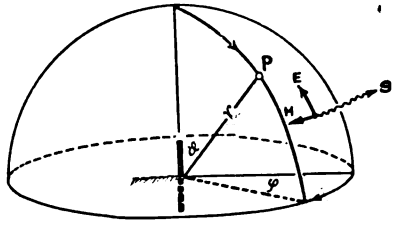

Figure 1—Spherical Coördinates

in which all electrical quantities (electric force (E), current density (i), resistance (R), etc.), are measured in electrostatic c. g. s. units, and the magnetic quantity (magnetic force (H)) in electro-

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magnetic units. The electric and magnetic forces at any point due to the current are found by first calculating the scalar and vector electromagnetic potentials ϕ and A by integration, and then deriving E and H by differentiation, as follows:

$$E = -Grad\,\phi - \frac{1}{c}\frac{\partial A}{\partial t}; \tag{1}$$

$$H = Curl\,A; \tag{2}$$

where

$$\phi = \int\int\int \frac{\rho\left(t \pm \frac{r}{c}\right)}{r}\,dv, \tag{3}$$

$$A = \int\int\int \frac{i\left(t \pm \frac{r}{c}\right)}{cr}\,dv. \tag{4}$$

E and H being found, the flow of energy, which we shall suppose to be represented by the Poynting-vector,[3] S, can be found; thus:

$$S = \frac{c}{4\pi}[E \times H.] \tag{5}$$

Since the earth is assumed to be perfectly conducting and there are no sinks of power in the field, the total power escaping from the antenna is found by integrating the normal component of the Poynting-vector over any surface inclosing it, it being a matter of indifference what surface is taken. For convenience we take the surface to be a hemisphere with center at the antenna base. The time-average of the total power is next found, and from this the resistance equivalent to radiation is derived by dividing it by the average current square. The essential steps of this process are given below; to save space all trifling algebraic details are omitted.

1. Assumed Current Distribution

The current amplitude is assumed to be sinusoidal as a function of position along the antenna wire (see Figure 2), and the instantaneous current is sinusoidal in time. Denoting the current amplitude at the base by I_o and the current amplitude at the "current-loop" by I, the instantaneous current at a point distant x from the earth is given by:

[3] It will be noted that the Poynting-vector gives precisely the same representation of the energy flow as Macdonald's energy-vector (Compare: "Electric Waves," page 72, Cambridge, 1902), because we are dealing with a periodic source and Macdonald's extra term, $\frac{d}{dt}[H \times A]$ vanishes when integrated over a complete period.

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$$i = I \sin \frac{2\pi(l-x)}{\lambda} \sin \omega t,$$

$$= I_o \frac{\sin \frac{2\pi(l-x)}{\lambda}}{\sin \frac{2\pi l}{\lambda}} \sin \omega t. \qquad (6)$$

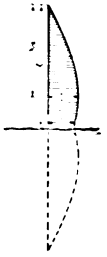

FIGURE 2—Assumed Sinusoidal Variation of Current Amplitude Below the Fundamental

It will be convenient to use the abbreviation: $a = \frac{2\pi l}{\lambda}$, or since we will take the fundamental wave length, λ_o, to be equal to the length of the wire multiplied by 4, $a = \frac{\pi \lambda_o}{2\lambda}$, λ being the operating wave length.

2. Calculation of the Vector Potential

The vector potential may be either of the accelerated or retarded variety and is obtained by integrating the current over the antenna and its image. Using the retarded potential:

$$A = \frac{I}{c} \int_o^l \sin \frac{2\pi(l-x)}{\lambda} \sin \omega \left(t - \frac{r'}{c}\right) \frac{dx}{r'} + \int Image. \qquad (7)$$

A is given in a point of the surface of the sphere for which r is constant. The distance from the integration element to this point is represented by r'. Now we will assume that the radius of this sphere is large compared with the length of the antenna. This permits us to neglect the effect of the variation of r' on the amplitude, but we can never neglect its effect on the phase, no matter how large r may be. We have approximately:

$$r' = r - x \cos \theta$$

The effect of the image is got by reversing the sign of $\cos \theta$. Performing the integration, we get:

$$A = -\frac{I \lambda}{\pi r c} \sin \omega \left(t - \frac{r}{c}\right) \frac{[\cos a - \cos(a \cos \theta)]}{\sin^2 \theta}, \qquad (8)$$

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which gives the retarded vector potential at a point the coordinates of which are r and θ.

3. Calculation of the Electric and Magnetic Field Intensities

It is unnecessary to calculate the scalar potential because we shall only require the electric force at a large distance from the antenna and in the self-conjugate field which exists at this distance, E and H are equal and orthogonal to each other and to the direction of propagation. Hence a calculation of H thru:

$$\boldsymbol{H} = Curl\, \boldsymbol{A}$$

suffices also for E. For the important components we have:

$$E_\theta = H_\phi = Curl_\phi A = \sin\theta \frac{\partial A}{\partial r} + \frac{\cos\theta}{r} \frac{\partial A}{\partial \theta}, \tag{9}$$

Performing the indicated operation and ignoring terms in r^{-2}:

$$E_\theta = H_\phi = \frac{2I}{cr} \cos\omega\left(t - \frac{r}{c}\right) \frac{\cos a - \cos(a\cos\theta)}{\sin\theta}. \tag{10}$$

E is entirely longitudinal and H is latitudinal.

4. Calculation of the Power Radiated

Since E and H arenormal to each other and to the spherical surface of integration, the Poynting-vector is entirely normal to the surface so we have simply to integrate the absolute value:

$$S = \frac{c}{4\pi} H_\phi^2, \tag{11}$$

over the hemisphere. The surface element is $ds = 2\pi r^2 \sin\theta d\theta$, hence:

$$\text{Total power} = \frac{2I^2}{c} \cos^2\omega\left(t - \frac{r}{c}\right) \int_o^{\frac{\pi}{2}} \frac{[\cos a - \cos(a\cos\theta)]^2}{\sin\theta} d\theta. \tag{12}$$

This integral apparently cannot be evaluated in simple functions. Pierce's method leads naturally to the same integral, which he attacks boldly, expending the integrand in series and integrating term by term. With the range of values of a with which Pierce was concerned (wave lengths *above* the fundamental) the convergent series so obtained can be computed with moderate patience and a calculating machine, but for wave lengths below the fundamental the argument becomes so large that the labor of calculation has appeared to me prohibitive. It is possible to develop the integral by successive partial integrations in an

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asymptotic series in inverse powers of a, and this is useful for large a, but there is still a region immediately below the fundamental where a is too small for accurate work. A lazy man thus finds himself between a convergent Scylla and a divergent Charybdis.

The simplest functions in terms of which the integral could be expressed seemed to be the following, the notation $S_1(x)$ and $S_2(x)$ being my own:

$$S_1(x) = \int_o^x \frac{1-\cos z \, . \, dz}{z} = \frac{x^2}{2 \, . \, 2!} - \frac{x^4}{4 \, . \, 4!} + \frac{x^6}{6 \, . \, 6!} - \frac{x^8}{8 \, . \, 8!} + \quad (13)$$

$$S_2(x) = \int_o^x \frac{\sin z \, dz}{z} = x - \frac{x^3}{3 \, . \, 3!} + \frac{x^5}{5 \, . \, 5!} - \frac{x^7}{7 \, . \, 7!} + \quad (14)$$

Being convinced, from the simplicity of these functions, that thay had important mathematical properties, and should be fundamental, I made a search of the literature and was gratified to find that 4-place tables of the Sine-integral, $Si(x)$ and the Cosine-integral, $Ci(x)$ were available[4] in terms of which S_1 and S_2 could be expressed as follows:

$$S_1(x) = \log x + \gamma - Ci(x) \quad (15)$$

$$S_2(x) = Si(x); \quad (16)$$

where γ is Euler's constant $= 0.5772$. Thus the laborious computation of the series (13) and (14) is avoided.

The integral for the power is expressed in terms of S_1 and S_2 as follows, after the time average has been taken:

$$\overline{\text{Total power}} = \frac{I^2}{c}\left[\cos^2 a \, S_1(2a) - \frac{1}{4}\cos 2a \, S_1(4a) - \frac{1}{2}\sin 2a \left(S_2(2a) - \frac{1}{2}S_2(4a)\right)\right]. \quad (17)$$

5. Expression for the Radiation Resistance

The radiation resistance is defined as the average power radiated per second divided by the average current square at the point of the antenna at which the power is introduced. If the resistance is calculated at the base we have to divide by $I_o^2/2$, I_o being the base current. Below the fundamental the current-loop (point at which the amplitude is maximum) moves up from the earth and at $\lambda_o = 2\lambda$ is situated half-way up the antenna. The resistance referred to this point is also of interest, as well as the usual base resistance; we shall represent it by R (loop), the

[4] Jahnke and Emde: "Funktionentafeln," page 19 (Teubner's reprint, Leipzig, 1923), or J. L. Glaisher: "Phil. Trans.," London, 160, page 368, (1870).

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base resistance being denoted by R_o. The complete expressions for the radiation resistance referred to these points follow:

$$R_o = \frac{2}{c \sin^2 a}\left[\cos^2 a\, S_1(2a) - \frac{1}{4}\cos 2a\, S_1(4a) - \frac{1}{2}\sin 2a\left(S_2(2a) - \frac{1}{2}S_2(4a)\right)\right]$$

. . . . electrostatic c. g. s. units,

$$= \frac{60}{\sin^2 a} \text{ [same bracketted expression] ohms} \quad (18)$$

$$R(loop) = 60 \text{ [same bracketted expression] ohms} \quad (19)$$

Formulas (18) *and* (19) *give the radiation resistance of a vertical single wire antenna over perfect earth for a sinusoidal current distribution,* R_o *being the resistance referred to the base, and* R *(loop) being the resistance referred to the point of the antenna at which the current-loop is situated in the case of wave lengths below the fundamental.* $a = \frac{\pi}{2}\frac{\lambda_o}{\lambda}$, *where* λ_o *= fundamental wave length and* λ *= the operating wave length.*

Computations, Tables, and Curves

For values of a up to about 4, the values of Si and Ci can be taken from the 4-place tables of Jahnke and Emde, a being chosen so that $4a$ is an integer (to avoid interpolation). Between $x = 16$ and $x = 20$, however, there is an important gap in the tables, corresponding to wave lengths between one-half and one-third the fundamental. This gap is filled in by computations from the following asymptotic series, the terms of which have a sufficiently rapid descent to make the remainder ignorable after one or two terms.

$$S_2(x) \backsim \frac{\pi}{2} - \frac{\cos x}{x}\left(1! - \frac{2!}{x^2} + \frac{4!}{x^4} - \ldots\right) - \frac{\sin x}{x}\left(\frac{1!}{x} - \frac{3!}{x^3} + \ldots\ldots\right), \quad (20)$$

$$S_1(x) \backsim \log x + .5772 - \frac{\sin x}{x}\left(1! - \frac{2!}{x^2} + \frac{4!}{x^4} - \ldots\right) + \frac{\cos x}{x}\left(\frac{1!}{x} - \frac{3!}{x^3} + \ldots\right) \quad (21)$$

With these formulas and tables, the computation easily proceeds, giving the results tabulated in Table I.

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TABLE I

Radiation resistance in ohms of a vertical antenna over perfectly conducting earth, the current distribution being assumed sinusoidal.

λ/λ_o	a	R_o (*base*)	R (*loop*)	λ/λ_o	a	R_o (*base*)	R (*loop*)
1.40	1.122	15.08		0.524	3.0	5270.	104.54
1.20	1.310	22.10		0.507	3.1	58200.	101.1
1.15	1.368	24.75		0.492	3.2	27900.	97.20
1.05	1.497	31.64		0.448	3.5	650.	80.0
1.00	1.570	36.54	36.54	0.418	3.75	191.	62.52
0.98	1.6	40.0	38.61	0.392	4.00	88.0	50.40
0.872	1.8	55.5	52.80	0.369	4.25	55.8	44.66
0.785	2.0	83.0	68.82	0.349	4.5	46.0	44.16
0.700	2.25	143.5	86.57	0.342	4.6	46.8	46.42
0.654	2.4	210.	95.45	0.314	5.0	79.5	73.01
0.571	2.75	730.	106.0				

For easy reference the results are also shown graphically in Figure 3, the dotted curve representing the resistance at the base, the full line curve representing the resistance at the current-loop.

I have carried thru the computations a short distance above the fundamental in order to make connections with Pierce's computations, and find a very satisfactory numerical agreement.

For wave lengths substantially above the fundamental, formula (18) is not convenient, since the resistance becomes increasingly small and is given by the difference between two functions of approaching equality. In this case a simplification is possible, as pointed out by Cutting[5], which follows from the fact that the current distribution approaches linearity as a decreases, and for small λ_o/λ we have very approximately:

$$R_o = 2.5\,\pi^2\left(\frac{\lambda_o}{\lambda}\right)^2 = 40\pi^2\left(\frac{l}{\lambda}\right)^2 \text{ ohms.} \qquad (22)$$

Returning to the curves, at first sight it seems odd that the maximum of resistance (loop resistance) should occur somewhat short of the half-wave length, at $\lambda/\lambda_o = 0.56$ rather than at 0.50; similarly that the minimum should fall short of the 1/3 position. This led me to check the formula and computations very carefully, but no error was revealed. Upon more mature consideration, there seems to be no reason why the curve should not be of this form.

The calculations are carried to a wave length one-third the fundamental. It might seem desirable to extend these calculations to the inverted "L" type of antenna, thus supplementing

[5] Fulton Cutting: PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, 10, page 129, (1922).

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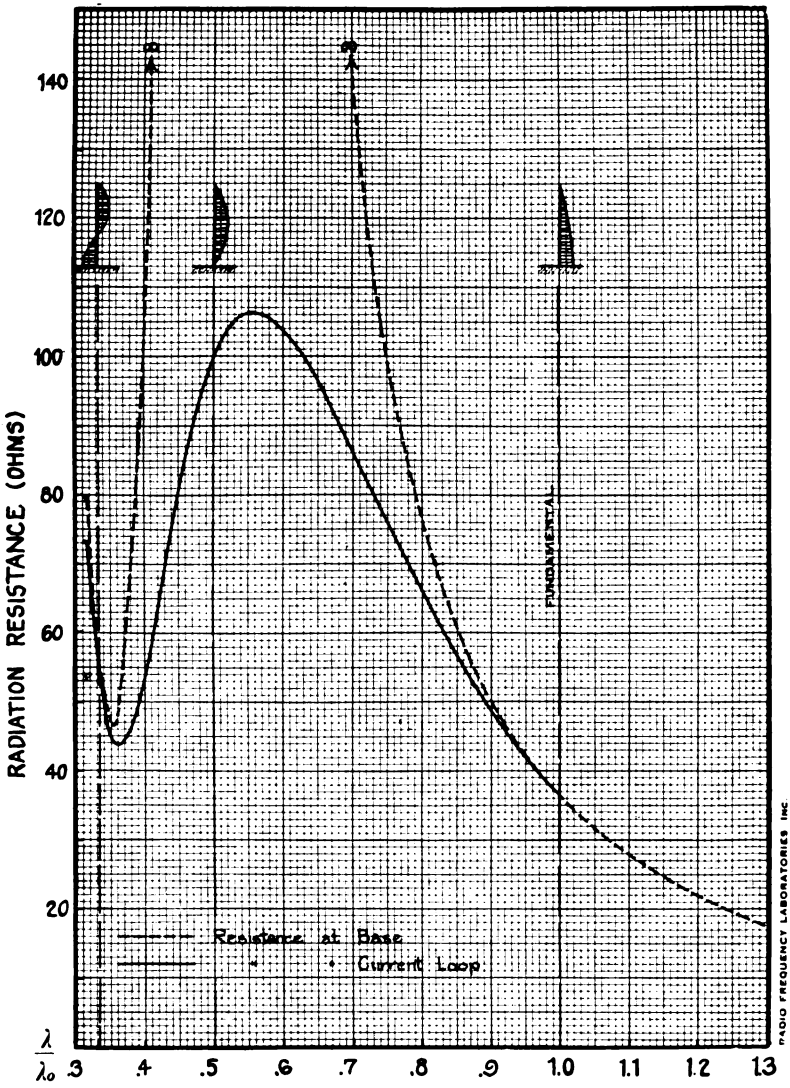

FIGURE 3—Radiation Resistance of Simple Vertical Antenna Over Perfect Earth for Wave Lengths Below the Fundamental

Pierce's study of the resistance of this type of antenna at wave lengths above the fundamental. But in view of the labor involved and the lack of faith which must be entertained for the fundamental assumptions—perfect earth and sinusoidal current distribution—I have not considered this to be worth while.

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(Note, added July 15, 1924:—Professor G. W. Pierce, commenting upon this paper, informs me that some calculations of the radiation resistance of a vertical antenna below the fundamental wave length were made in 1919 by Dr. Yu Ching Wen, a former pupil of his. This work exists in the form of a thesis entitled "Theoretical Treatment of the Radiation Resistance of Antennas," in the Library of Harvard University, and has not been published. Mr. Robert F. Field, of Cambridge, has kindly consulted this thesis for me and sends the following values for the resistance at the antenna base calculated by Wen:

λ/λ_0	1.00	0.94	0.90	0.82	0.80	0.76	0.70	0.66	0.64
R	36.6	43.7	50.3	71	78	96	162	212	236

These calculations go down to 0.64 λ_0 and agree satisfactorily with my values (dotted curve, Figure 3).)

Cambridge, Massachusetts,
November 8, 1923.

SUMMARY: In this paper, the radiation resistance of a filamentary vertical antenna erected over perfectly conducting plane earth is calculated for wave lengths down to 0.3 of the fundamental. Pierce's assumption of sinusoidal variation of the current amplitude along the antenna forms the basis of an application of the methods of the electron theory, whereby the e.m. potentials are found by integration and the field and Poynting vectors subsequently derived by differentiation. An improved method of computation, using the sine- and cosine-integrals, is introduced. The numerical results are tabu-'ated and exhibited graphically.

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ON THE OPTIMUM TRANSMITTING WAVE LENGTH FOR A VERTICAL ANTENNA OVER PERFECT EARTH*

By

Stuart Ballantine

(John Tyndall Scholar in Physics, Harvard University, Cambridge Massachusetts)

The question as to the best operating wave length for a transmitting antenna is of considerable theoretical and practical interest, and one which does not appear to have received much published discussion. The mathematical difficulties of the theoretical study are very formidable if the actual conditions of the practice are imposed and an effort made to take into account all kinds of antenna forms and grounding methods, and the variations of earth conductivity and susceptibility. But certain simple cases can be advantageously treated mathematically, leaving the more complicated and large scale engineering projects to be tested by model experiments, the model being constructed in accordance with the principles of dimensional similarity. In a forthcoming series of papers I propose to publish the results of theoretical and experimental work on this subject, and in this preliminary article shall consider the simplest possible case, namely, that of a simple vertical antenna over a perfectly conducting plane earth.

We shall start out with the additional assumption that there are no dielectric or other losses present, so that the antenna resistance will be entirely due to radiation. In a companion paper entitled "The Radiation Resistance of a Simple Vertical Antenna at Wave Lengths Below the Fundamental," published in these Proceedings, I have discussed the radiation from such an antenna in these circumstances, and calculated the radiation resistance for wave lengths down to 0.3 of the fundamental. The formulas and results of this investigation form the basis of the present inquiry.

For terrestrial point-to-point communication the function of a transmitting antenna is to produce an electric force at a receiving station located on the horizon, therefore the best operating

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wave length is that for which the electric force on the horizon is greatest for a given power input into the antenna. There is a considerable confusion of ideas on this point due to an excessive use of the conception of radiation resistance, the temptation being to regard the point at which the radiation resistance is highest as the best wave length. But since the object is to heat the wires of the receiving station, and not the universe, it is easy to see that the best wave length is that which will cause the antenna to radiate most of its energy at a low angle and along the earth's surface. Unless communication with aircraft is desired, or the efficiency of the Heaviside layer is more definitely established, there seems to be no point in radiating the antenna's energy up into the air.

2. Distribution of the Radiant Energy—The energy beam diverges from the power source at the base of the antenna, or if the power is not applied at the base, from the point of its application. There is no divergence from the antenna wires themselves, the energy being merely reflected by them. No detailed inquiry of the distribution of energy at points near the antenna will be made; the zone of chief interest is the so-called "wave zone" which exists at some distance from the antenna and in which the wave is self-conjugate, the electric and magnetic forces being equal. The intensity of the energy beam, or the rate of flow of the energy in the wave, is generally and conveniently represented by the Poynting vector, the magnitude of which is:

$$S = \frac{c}{4\pi} E \times H, \tag{1}$$

where c is the light-velocity, E the electric force, H the magnetic force (both in Gaussian units). The direction of this vector is at any instant at right angles to both E and H, so that at a reasonably large distance from the antenna, where E and H lie in the surface of a sphere surrounding it, S is normal to the surface and therefore directed away from the antenna. Also E and H being equal in the wave-zone, S can be simply written as:

$$S = \frac{c}{4\pi} E^2 = \frac{c}{4\pi} H^2 \tag{2}$$

In the paper just cited will be found formulas for calculating E at various angles, θ, from the pole, for different wave lengths. From these S can be calculated and plotted as a function of the polar angle in the form of a *distribution diagram*. Such a diagram will be of interest in showing how much energy is radiated in the

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various angular directions for different current distributions in the transmitting antenna, the current distribution being governed by the wave length and the height of the antenna wire. Several diagrams of this sort will be given later.

3. Calculation of the Optimum Transmitting Wave Length—As previously remarked the signal-producing value of the antenna's radiation is represented by the electric field intensity E on the horizon. On the horizon, θ of formula (10) of the previous paper is $\pi/2$ so that the amplitude of E is given by:

$$E=\frac{2I}{cr}\left[1-cos\left(\frac{\pi\lambda_o}{2\lambda}\right)\right], \tag{3}$$

where I is the current at the current-loop of the antenna, λ is the operating wave length and λ_o is the fundamental wave length. Since the antenna resistance is entirely radiational, R_r:

$$I^2R_r=\text{antenna power}=P, \tag{4}$$

R_r being the radiation resistance at the current-loop, and

$$I=\sqrt{\frac{P}{R_r}}, \tag{5}$$

What we are interested in making a maximum is the ratio $E/\sqrt{P}$, which is given by substituting (5) in (3):

$$\frac{E}{\sqrt{P}}=const.\frac{1-cos\left(\frac{\pi\lambda_o}{2\lambda}\right)}{\sqrt{R_r}} \tag{6}$$

The data for R_r as a function of $a=\frac{\pi\lambda_o}{2\lambda}$ is given in the companion paper (previous citation) for a vertical antenna over perfect earth. Using these results, the expression (6) has been calculated and plotted in Figure 1; the full-line curve represents $E/\sqrt{P}$, the dotted-line curve the radiation resistance (loop). The best operating wave length for this idealized antenna is thus seen to be closely equal to 0.39 of the fundamental. A comparison with the R_r curve convinces us that no deductions of any value can be drawn from the variation of radiation resistance alone.

4. Distribution Diagrams—In Figures 2 and 3 are plotted the diagrams showing the distribution of radiant energy at various altitudes for several important wave lengths. In all cases the input power to the antenna is the same. These are polar graphs and the strength of the radiation (Poynting vector) is given as the distance from the antenna base to the curve for a particular

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value of the polar angle, θ. In all cases it will be noticed that the usual "dead spot" occurs directly over the antenna.

In the first diagram, Figure 2a, for wave lengths very much greater than the fundamental, altho the main tendency is for the energy to flow along the earth, a considerable amount is radiated at high angles; the 45° radiation, for example, is about half as intense as that along the earth. As the wave length is decreased to the fundamental and beyond, more and more of the energy is radiated along the earth, and less gets up into the air.

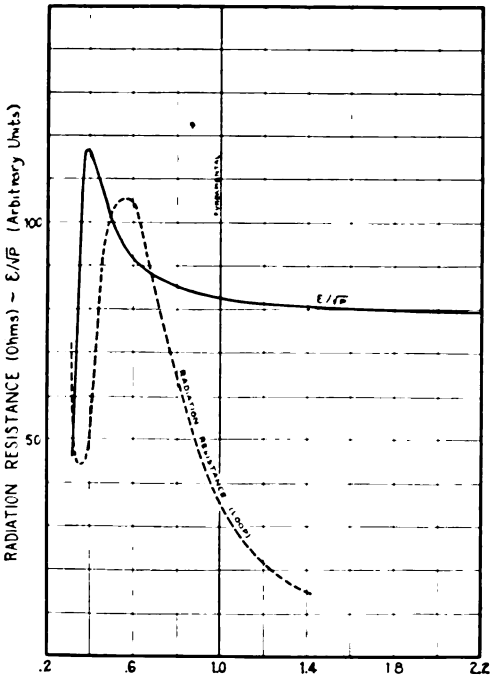

FIGURE 1—Vertical Antenna Over Perfect Earth; Variation of Electric Force on the Horizon for Given Power Input as a Function of the Ratio: Operating Wave Length λ Fundamental Wave Length λ_0. Best Transmitting Wave Length = $0.39\,\lambda_0$.

When the operating wave length is decreased below the half-fundamental wave length, the wave breaks up into two parts, and a secondary wave, complete in itself and unguided by the earth, appears at high angles. These waves are separated by a conical surface, $\theta = const.$, in which the electric and magnetic forces are zero. There are then two points of maximum radiation. This

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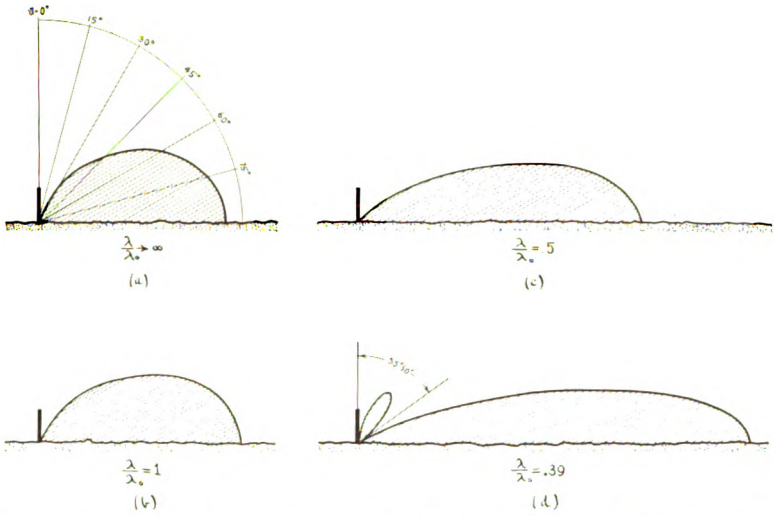

FIGURE 2—Vertical Antenna Over Perfect Earth; Diagrams Representing Intensity of the Radiation (Poynting Vector) at Various Angles of Altitude for Various Wave Lengths

is illustrated in Figure 2 (d), which represents the case $\lambda/\lambda_o = 0.39$, which as was shown in a previous paragraph is the best operating wave length. Here the radiation along the earth in the direction of the distant receiving station is very intense, and relatively little energy is wasted in high angle radiation. The maximum of the secondary wave occurs at $\theta = 35°$, and $\theta = 55°50'$ is the conical surface of zero radiation. As the wave length is shortened below this point, the primary radiation decreases very rapidly, until at $\lambda/\lambda_o = 1/4$ the separating cone becomes the earth surface and the primary wave has been completely supplanted by the secondary wave. This case is shown in Figure 3, and is of special interest.

5. SPECIAL TRANSMITTING ARRANGEMENT FOR TESTING UPPER-ATMOSPHERIC REFLECTION—In most of the experimental work that has been done to establish or disprove the theory of the Heaviside-layer and upper-atmospheric reflection, ordinary transmitting arrangements have been employed. At the receiving point it is therefore necessary to provide an elaborate system of loops to separate the direct ray along the earth from the ray reaching the receiver by reflection from the upper atmosphere. The ingenious experimental arrangements described by Eckersley ("Radio Review") are of this nature. Several years ago it occurred to me that the experiment would be very conveniently

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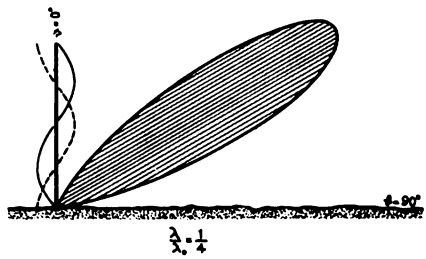

FIGURE 3—Distribution Diagram for the Case: Height = Wave Length, in which the Radiation Along the Earth is Zero. Same Scale and Power Input as Figure 2

simplified if the transmitted wave could be entirely directed up into the air, thus eliminating the direct ray along the earth. The transmitting arrangement for high angle radiation could consist of a vertical antenna erected preferably over sea water and excited at a wave length one-quarter of its fundamental. At this wave length the voltage and current distribution would be approximately that shown in Figure 3, so that the radiation resistance being high (at the base) the antenna would naturally be excited by "voltage excitation," that is, it would be connected in parallel with an anti-resonant $L\,C$ circuit. The distribution of energy about such a radiator is shown in Figure 3. The radiation centers about an altitude of $\theta = 57°$, and if the conductivity at the base be sufficiently good there will be no disturbing direct ray along the earth. The possibility would then be that any radiation reaching a distant receiving station would be due to reflection and refraction in the upper-atmospheric layers. The experiment would preferably be conducted entirely over sea-water on account of its superior conductivity, because, as I shall show in another paper, one of the effects of imperfect earth is to bring down to the earth energy originally sent up into the air. The elimination of the direct ray would permit an easy study of the polarization and other characteristics of the reflected ray. As to the practical aspects, the height of the antenna would have to be equal to the wave length so that a comparatively short wave length will be imposed.[1] This in itself will not be disadvantage-

[1] The necessity for an antenna length actually equal to the wave length can be obviated without changing the required current distribution by decreasing the velocity of the wave along the wire by artificially loading with inductance coils, or by constructing a helical antenna of not too large diameter. Two inductance coils, situated ¼ and ¾ of the distance from the earth to the end of the antenna, respectively, and kept approximately equal in value, will suffice to produce the required double current-loop distribution in a short antenna. Oscillating in this mode the antenna can be looked upon as consist-

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ous, provided proper audions are used for receiving, and with shorter waves, smaller reflecting masses can be detected. A wave length of 50 meters is feasible and demands an antenna height of 154 feet (50 m.).[1] Being forced to abandon this work on account of failing health, I have included this description of the plan in the hope that it might interest other investigators; it offers a very promising method of attacking this fascinating puzzle.

November 8, 1923.

SUMMARY: This paper considers the idealized problem of the transmission from a perfect vertical antenna over perfectly conducting plane earth. The distribution of current amplitude along the antenna is assumed to be sinusoidal. The antenna resistance is entirely radiational and has been calculated in a companion paper. The amount of energy radiated in various altitudinal directions is calculated, and distribution diagrams are given illustrating special cases. The conditions for the greatest economy of radiated energy are formulated for the case of a terrestrially located receiving station, the optimum wave length being determined as 0.39 of the fundamental. A transmitting arrangement of special interest for Heaviside-layer experiments is also described.

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ing of two Hertzian antennas in series, each with its inductance load in the middle. One Hertzian antenna is positive, the other negative; the radiation effects cancel along the equatorial plans and on account of the phase change do not completely cancel at other angles. The radiation from the artificially loaded antenna will be less, of course, than from the natural antenna of Figure 3.

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ELECTRICAL CONSTANTS OF DIELECTRICS FOR RADIO FREQUENCY CURRENTS*

By

R. V. Guthrie, Jr.

(State University of Iowa, Iowa City)

I. Introduction

The electrical constants of dielectrics are the dielectric constant K and the power factor ψ, where ψ is the power factor of a condenser using the given material as a dielectric. Determinations of power factors for a few materials have been made by MacLeod,[1] Bryan[2] and others. Tables of dielectric constants are to be found in engineering handbooks, "Smithsonian Physical Tables," and elsewhere, but these were made for the most part using currents of zero frequency. Therefore it seems of interest to determine the dielectric constants and power factors of a number of insulating materials over a range of radio frequencies.

The method employed was the obvious one and consisted of measurements of the capacity and resistance of a carefully made parallel-plate condenser with air and the material in turn between the plates. From these measurements the dielectric constant and the power factor were calculated.

Consider the arrangement in Figure 1 consisting of a condenser C' and a resistance r'' in series. The impedance of the combination from A to B is given by

$$Z_{AB} = \frac{1}{j\,\omega\,C'} + r'' \qquad (1)$$

where $j = \sqrt{-1}$, and $\omega = 2\pi$ times the frequency, all quantities being expressed in electrostatic units. From a consideration of Figure 1, it is seen that $E_{r''}$ across the resistance is in phase with the current I, and the $E_{C'}$ across the condenser is 90° behind I. From Figure 2 we see that the resulting emf. is less than 90° out of phase with the current by the phase difference ψ, the

*Received by the Editor, June 10, 1924.

[1] MacLeod, "Phys. Rev.," volume 21, January, 1923, page 53.

[2] Bryan, "Phys. Rev.," volume 22, October, 1923, page 399.

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power factor being given by $sin\ \psi$, but for small angles the sine is the same as the tangent, then

$$\text{Power factor} = tan\ \psi = r''\ \omega C' \tag{2}$$

FIGURE 1

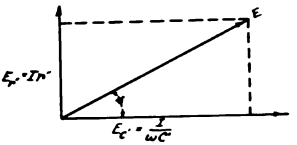

FIGURE 2

If, however, C' be measured by substitution against a standard condenser, not the real capacity but the apparent or effective capacity will be obtained. The equivalent of Figure 1 may be represented as in Figure 3 by a pure capacity C with a resistance r' in parallel. Here the impedance is given by

$$Z_{AB} = \frac{1}{j\omega C + \frac{1}{r'}} \tag{3}$$

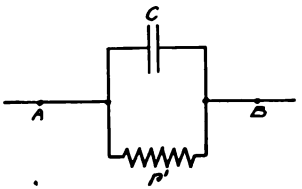

FIGURE 3

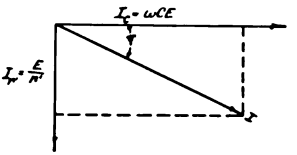

FIGURE 4

It may be seen that the current thru r' is 90° out of phase with that thru C. The resultant current leads the impressed emf. by an angle which is less than 90° by the phase difference ψ. The power factor is again the sine of ψ, which for small angles is the same as the tangent of ψ, so that

$$\text{Power factor} = tan\ \psi = \frac{1}{r'\ \omega C} \tag{4}$$

The source of power losses in a condenser may be represented by a series resistance as in Figure 1, or by a parallel resistance as in Figure 3, the power factors for small angles being the same as the phase differences as given by equation (2) or (4) in which, for small angles, the tangents may be taken as the angles. Before the application of these formulas it is necessary

[3] "Circular of Bureau of Standards," Number 74, "Radio Instruments and Measurements," page 180.

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to determine the appropriate values of the capacity, resistance, and ω.

By including AB in a suitable measuring circuit to be described, r'' may be determined by resistance variation,[3] and C' by substitution against a standard condenser, for any value of ω. To obtain a relation between the measurable quantities C' and r'', and the quantities C and r' we combine equations (1) and (3), separate the real parts from the imaginary ones in both members of the equation, equate reals to reals and imaginaries to imaginaries, and solve the two resulting equations simultaneously. We obtain

$$C = \frac{C'(r'-r'')}{r'} \tag{5}$$

$$r' = \frac{1}{C'^2 r''^2 \omega} + r'' \tag{6}$$

Since C' and ω are known from measurements, r' and C are determined by equations (5) and (6), therefore we have the quantities r', r'', C', C, and ω all measurable or calculable.

II. Method and Apparatus

The schematic diagram of apparatus used is shown in Figure 5, the portions to the right of the dotted line being the coil of a thermionic tube oscillator for the generation of continuous waves, and that to the left of the line being the measuring circuit. The measuring circuit consisted of a coil L_2, the terminals of which dropped into the mercury cups a and b, thus allowing rapid change to coils of any desired inductance; also a capacity C_1 or C_2 depending on which way the mercury switch $c\,d\,e$ was thrown. There was included in this circuit a thermocouple T which actuated the galvanometer G, and two mercury cups f and g, between which standard resistance links r of various values might be connected. C_1 is a standard quartz-insulated condenser, which was assumed to be perfect and was carefully calibrated.

A photograph of condenser C_2 is shown in Figure 6. It had two plates m and n, 7.3 cm. in diameter, m being insulated from the other parts by the quartz rod q. The similar plate n was capable of vertical motion by a screw s. The dielectric to be studied was placed between these plates, thus forming a condenser having a solid dielectric. The position of n at any time could be determined by the dial d and the vernier v.

The insulating material was placed between the plates of C_2, which was connected into the circuit by joining c and d by a heavy copper link between the cups. The condenser C_2 could

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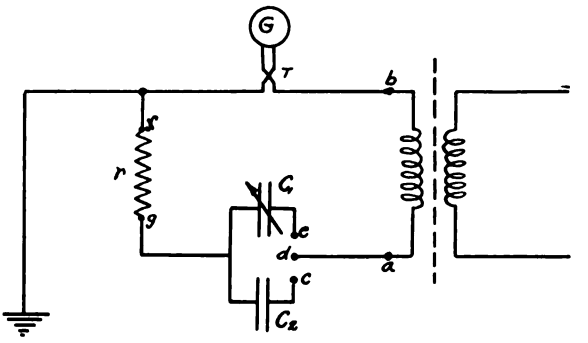

FIGURE 5

then be represented as in Figure 1. Resonance was effected in the L_2 C_2 circuit by varying the frequency of the input, all measurements being made at resonance. The resistance of the circuit L_2 C_2 was then determined by resistance variation.

When a link of negligible resistance was connected between g and f, suppose a beam of light reflected from a small mirror on the coil of the galvanometer G, read a deflection d_1 on a scale not shown in the diagram. When the link is replaced by a standard resistance of R_1 ohms, the deflection would decrease to some value d_2, then the resistance R_x of the circuit is given by

$$R_x = \frac{R_1}{\sqrt{\frac{d_1}{d_2} - 1}} \tag{7}$$

Without changing the frequency of the input, the standard C_1 was substituted for C_2 by changing the connection $d\,c$ to $d\,e$, adjustments to resonance being made with C_1. If we represent the circuit resistance as R_x', then $R_x' - R_x$ is the equivalent series resistance of the condenser C_2 with the solid dielectric and is r'' of equation (6). The capacity as read from the dial of C_1 is the pure capacity of C_1, but is the effective capacity of C_2 corresponding to C' of equation (6).

The wave length λ was determined by use of a Kolster decremeter calibrated by the Bureau of Standards. From the relation $\omega = 2\pi f = (2\pi \times 3 \times 10^8)/\lambda$, ω was determined.

Having now measured r'' and C', r' may be obtained from equation (6), then C is given by equation (5). After all quantities have been reduced to electrostatic units, the power factor is given either by equation (2) or (4), both giving the same result.

Knowing the pure capacity of the two-plate condenser with

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the dielectric, and also its capacity without the dielectric, the quotient of the former by the latter is, by definition, the dielectric constant. C has been calculated from equation (5), and the capacity without the dielectric must be obtained by substitution against the standard.

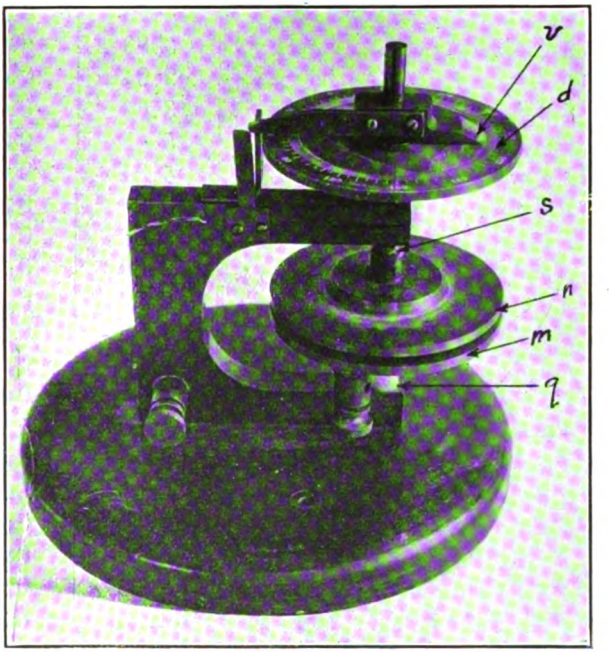

FIGURE 6

III. RESULTS

The constants of the materials investigated are given in the table. The first six were tested over a band of frequencies ranging from 600,000 to 1,700,000 cycles per second. It was found that the error introduced by assuming $C = C'$ was never more than 0.25 percent, which was much below the error of observation. The constants given are accurate to within about 2 percent. The dielectric constants were found to be nearly constant within the frequency limits mentioned, and the power factors changed gradually but according to no apparent rule.

If $C = C'$, then it is not necessary to use either equation (5) or (6). The dielectric constant is determined by using C' instead of C, and the power factor is determined by use of equation (2). Moreover, in applying (2), C' may be expressed in microfarads if r'' is expressed in ohms, thus eliminating the necessity

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TABLE OF CONSTANTS FOR FREQUENCY OF 10^6 CYCLES PER SECOND

Number	Material	Dielectric Constant	Power Factor
1	Hard Rubber	2.53	0.007
2	Celluloid	4.10	0.042
3	Formica	3.83	0.050
4	Mica	2.94	0.0004
5	Sulphur	3.03	0.006
6G	Glass	5.65	0.008
7	Petrite	5.34	>0.100
8	Bakelite Dilecto Gr. 20	3.86	0.057
9	Bakelite Dilecto Gr. 20	3.86	0.059
10	Hard Rubber	2.60	0.012
11	Vulcanized Hard Rubber	2.92	0.007
12	Hard Fibre (gray)	6.05	0.070
13	Hard Fibre (black)	5.26	0.052
14	Bakelite Dilecto	3.91	0.059
15	Formica (grade M)	3.60	0.051
16	Radion (black)	3.22	0.017
17	Hard Rubber (2 XX)	2.66	0.011
18	Hard Rubber (40)	2.57	0.011
19	Hard Rubber (35 R)	2.79	0.008
20	Hard Rubber (7A)	2.56	0.014
21	Red Fibre	4.35	0.054
22	Bakelite	3.87	0.037
23	Leatheroid	4.27	0.048
25	Mahogany	4.43	0.051
26	Celeron	4.12	0.049
27	Celeron	4.40	0.065
28	Celeron	3.73	0.057
29	Celeron	3.93	0.057
30	Fibroc	3.99	0.041
31	Bakelite No. 1	3.83	0.039
32	Bakelite No. 2	3.73	0.039
33	Bakelite No. 3	3.81	0.042
34	Bakelite No. 4	3.88	0.043
35	Bakelite No. 5	3.84	0.044
36	Bakelite No. 6	3.76	0.045
37	Vulcabeston No. 120	3.42	0.037
38	Vulcabeston No. 701	3.71	0.046
39	Vulcabeston No. 501	3.95	0.050

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of reduction to and from electrostatic units. This very materially lessens the labor of measurements and is the method followed in the determinations of the constants of the materials from number 6 to the close of the table.

For all materials in the table, the given constants are for currents of a frequency of 1,000,000 cycles per second, which corresponds to a wave length of 300 meters.

Measurements of the variation of resistance of coils with diameter of coil, pitch of winding, and size of wire, are in progress, and a report is expected to be submitted in the near future.

In conclusion the writer wishes to express his thanks to Dr. E. O. Hulburt for suggesting the problem and rendering frequent assistance during the experiment.

State University of Iowa.

SUMMARY: The source of power losses in a condenser may be considered as a resistance in series or in parallel with the condenser. If the resistance is considered as in series, the capacity must be considered as the effective capacity, and the power factor is given by a simple formula. If the resistance is considered as in parallel, the capacity must be considered as a pure capacity, and knowing the pure capacity with the given dielectric, and with air in turn between the plates of the condenser, the quotient of the former by the latter is the dielectric constant. The dielectric constants of several solid dielectrics were found to be approximately constant over the band of frequencies investigated, and the power factors changed according to no apparent law. A table of electrical constants for a number of materials and for a frequency of 1,000,000 cycles per second is given.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

ISSUED AUGUST 26, 1924–OCTOBER 28, 1924

BY

JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, D. C.)

1,499,979—C. M. Gray, filed August 15, 1921, issued July 1, 1924.

TUNING DEVICE FOR RADIO RECEIVING CIRCUITS, in which a fixed coil is provided with a second coil arranged in inductive relation to the fixed coil and mounted for rotary movement. A third coil is rotatively mounted in inductive relation to the aforementioned coils and has relatively few turns as compared with the second coil. The third coil is connected in series with the second coil and all coils are arranged for connection in the circuits of the system to be tuned.

1,506,046—W. H. Bullock, filed January 23, 1923, issued August 26, 1924. Assignor of 49 percent to Frederic W. Proctor, of New York.

RADIO RECEIVING SYSTEM, in which a plurality of signals transmitted on different wave lengths may be received by using the same antenna system without interference of one receiving circuit with respect to the other. An energy-absorbing circuit is employed in each of the receiving branches which prevents the flow of current in one direction for preventing the grid of one electron tube connected in circuit therewith from acquiring a static negative charge. The resistance of the absorbing circuit is too high to permit radio frequency oscillations, thereby preventing reaction of one circuit upon another while freely passing the desired signal frequency.

1,506,358—Otto Von Bronk, filed September 3, 1921, issued August 26, 1924. Assigned to Gesellschaft für drahtlose Telegraphie m. b. H. Hallesche, of Berlin, Germany.

SENDING ARRANGEMENT FOR RADIO TELEGRAPHY, in which the desired phase relations of the transmitted energy is regulated

* Received by the Editor, November 17, 1924.

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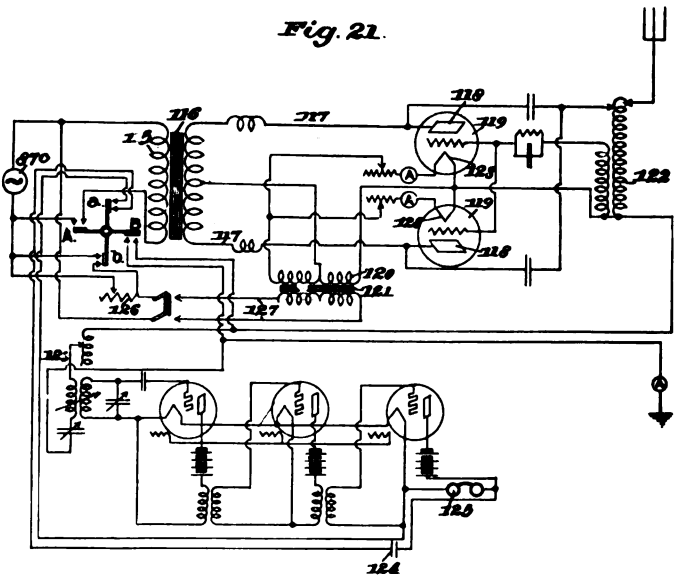
by arranging a plurality of separated radiators in positions to radiate currents of low intensity having a frequency difference equal to the frequency to be transmitted. These currents are received at each of the radiators and are employed to control the radiation from the radiators in proper phase relation.

1,506,468—W. C. White, filed October 28, 1920, issued August 26, 1924. Assigned to General Electric Company, of New York.

ELECTRON DISCHARGE DEVICE of the "soft" tube variety designed to be fully as sensitive as the best gas detectors formerly constructed, and as regular and reliable in its operation as the vacuum detector. The detector embodying the invention contains certain gases immune to the "clean-up effect," particularly the gases of the "rare" or noble gas group, such as argon, helium, neon, and even nitrogen.

1,506,580—H. E. Hallborg and H. R. Miller, filed October 25, 1920, issued August 26, 1924.

Fig. 21.



NUMBER 1,506,580—Quick Acting Amplifying Break Key for Radio Telegraphy

QUICK ACTING AMPLIFYING BREAK KEY FOR RADIO TELEGRAPHY for placing the receiver in condition for operation between the transmission periods. A quick acting break key is provided for controlling the connection of a shunting condenser across the

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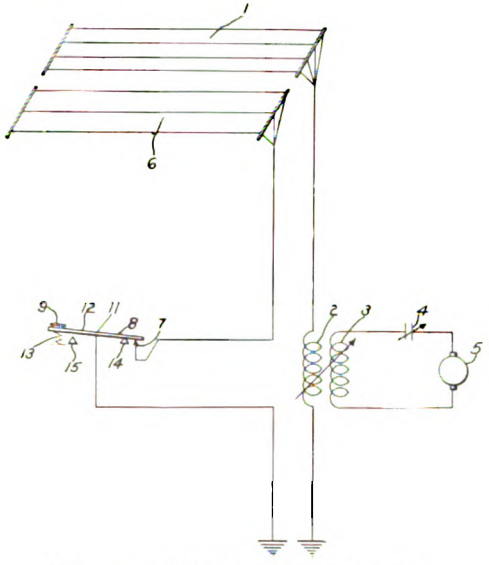
telephone receivers in the electron tube receiving system intermediate the times that the antenna is radiating the transmission signals. The shunting condenser prevents objectionable noise in the telephone system by reason of the radiating energy.

1,506,486—A. W. Hull, filed March 1, 1920, issued August 26, 1924. Assigned to General Electric Company, of New York.

Electric Oscillator, including an evacuated vessel with the electrodes arranged in such manner that the device will efficiently operate as an oscillator without making the grid positive. The cathode and anode are placed close to each other and control electrode or grid is so placed as to exercise a weak electrostatic control on one side of the cathode opposite the anode. The potential of the control member varies between the potential of the cathode and a potential which is negative with respect to the cathode.

1,506,736—L. Dorfman, filed August 3, 1921, issued September 2, 1924. Assigned to Westinghouse Electric and Manufacturing Company, of Pennsylvania.



Number 1,506,736—Signaling System

Signaling System having an antenna system consisting of two sections. A source of undamped radio frequency energy is connected to one of the antenna sections and a signaling key pro-

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vided in a circuit connecting the other antenna section. The antennas are electrostatically coupled and the frequency of the source of radio energy is such that its antenna is resonant when the signaling key is open. In this way radiation may occur to represent the signal characters while the energy is absorbed by one of the antenna sections between the signal characters.

1,506,742—C. Le G. Fortescue, filed June 25, 1921, issued September 2, 1924. Assigned to Westinghouse Electric and Manufacturing Company, of Pennsylvania.

MODULATING SYSTEM for radio transmitting systems in which the load on the generator is maintained substantially constant in either the up position or the down position of the telegraph key. A tuned antenna circuit having an inductance coil therein is provided, and located in inductive relation to the inductance coil is a controller primary inductance coil. A balancing inductance coil is connected in parallel with the controller primary coil thru contacts controlled by a telegraph key. A source of radio-frequency current is connected to the middle points of the controller primary and balancing inductance coils, and the key controls the flow of current thru said coils. Means may also be provided for maintaining the load on the radio-frequency source substantially constant. By reason of the coupling of the antenna inductance coil to the controller primary inductance coil, when the key is in the up position, no effective power is delivered through the transformer coupling, and the frequency of oscillation of the antenna circuit is greatly reduced, as compared with the operating frequency. This takes place by reason of the fact that, in the down position of the key, when the transformer is delivering power to the antenna, the primary coil of the transformer coupling is, in effect, short-circuited by the radio-frequency source.

1,506,781—J. E. Shrader, filed August 18, 1921, issued September 2, 1924. Assigned to Westinghouse Electric and Manufacturing Company, of Pennsylvania.

VARIABLE LEAKY CONDENSER, comprising a plurality of conductors and dielectric material treated with a conducting material interposed between the conductors. The distance between the conductors is made variable by a pressure varying screw bearing upon one of the conductors whereby the distance between the conductors is varied at the same time that the resistance may be increased or decreased. The spacing member may be of re-

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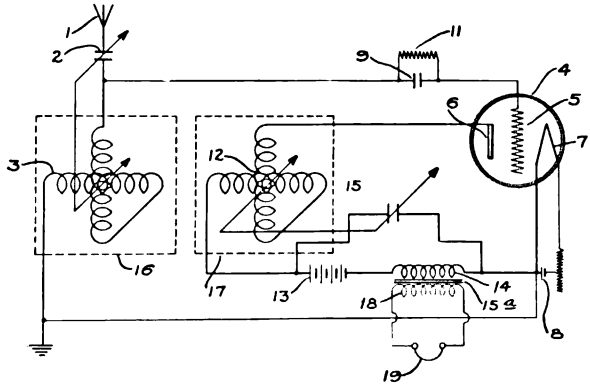
silient corrugated material adapted to be flattened under variable pressure. The dielectric material may be impregnated with India ink, forming a partial conductor. In this way a grid leak and grid condenser may be embodied in the same instrument.

1,506,799—P. E. Wiggin, filed July 9, 1921, issued September 2, 1924. Assigned to Westinghouse Electric and Manufacturing Company, of Pennsylvania.



Number 1,506,799—Radio Tuning System

Radio Tuning System for operation with an oscillatory or regenerative system including an electron tube. The input circuit of the tube is tuned by controlling the impedance of the circuit. The output circuit of the tube is tuned by simultaneously varying the impedance and the capacity of a variable by-pass condenser in the said circuit. The circuit is described with particular application to a radio receiver for the reception of broadcast signals.

1,507,016—L. De Forest, filed September 23, 1915, issued September 2, 1924. Assignor to De Forest Radio Telephone and Telegraph Company, of New York.

Radio Signaling System, employing an electron tube circuit arranged to function as an oscillator with an inductive coupling between the oscillation circuit and an antenna system. The oscillating system is employed as a generator of energy at a radio transmission system.

1,507,017 L. De Forest, filed March 20, 1914, issued September 2, 1924. Assigned to De Forest Radio Telephone and Telegraph Company, of New York.

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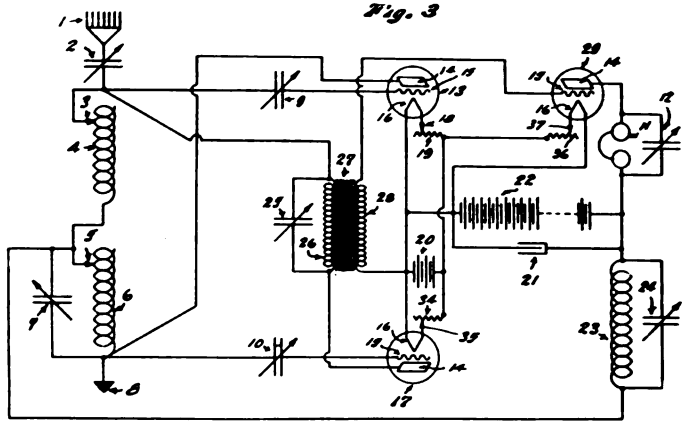
RADIO TELEPHONE AND TELEGRAPH SYSTEM, in which an electron tube circuit is employed, connected as an oscillator for generating oscillators at a transmitting station. The claims of this patent recite the tube structure and electrodes therein with circuits connecting the electrodes associated to react upon each other. The diagrams in the patent show the tube functioning as an oscillator at the radio transmitter.

1,507,689—F. G. Simpson, filed November 4, 1922, issued September 9, 1924.



NUMBER 1,507,689—Radio Telegraph and Telephone Receiving System

RADIO TELEGRAPH AND TELEPHONE RECEIVING SYSTEM, employing a local generator at the receiving station. A pair of valves are connected in the receiving circuit and arranged to permit the generator to supply uni-directional current impulses to the oscillator antenna circuit. The circuit including said generator is adjusted to be resonant at substantially twice the frequency of the oscillatory antenna circuit.

1,507,725—J. Bethenod et al, filed August 19, 1921, issued September 9, 1924.

DOUBLY TUNED RADIO RECEIVER, including one circuit tuned to a frequency greater than the signal frequency, and a second circuit tuned to a frequency lower than the signal frequency. The circuits are so arranged that the current therein may be opposed to prevent the response of the system to sudden discharges, such as static, but to permit the system to operate on the receipt of signals.

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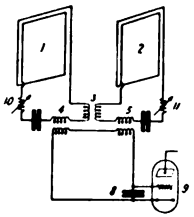
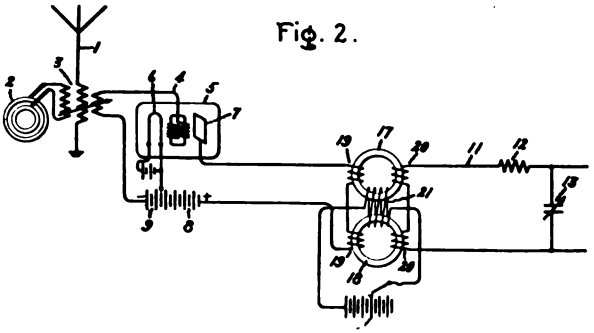
Fig. 2



NUMBER 1,507,725—Doubly Tuned Radio Receiver

1,508,151—E. F. W. Alexanderson, filed April 19, 1916, issued September 9, 1924. Assigned to General Electric Company, of New York.

Fig. 2.



NUMBER 1,508,151—Radio Signaling System

RADIO SIGNALING SYSTEM, for the reception of signals without disturbances from static effects. The patentee points out that in order to receive signals thru strays, a circuit must be provided to discriminate between these two sources of energy, by comparing the character of the energy over relatively long periods. A circuit is shown for prolonging the time during which the signal can act with accumulative effect on the receiving circuit considerably beyond the time during which disturbances can act. In order to prolong the time during which the signals and the disturbances may be compared, a tuned circuit of lower than radio frequency, but above audibility, is provided. The desired lowered frequency is obtained by means of interference beats between the incoming signals and the local source to produce a current of radio frequency having amplitude pulsations of much lower frequency, but above audibility. This current is used to produce an alternating current having a frequency corresponding to the frequency of the amplitude pulsations. A

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resonant circuit is provided for accumulating the energy of this alternating current and a detector is operated by the energy thus accumulated.

1,508,356—W. C. White, filed February 18, 1920, issued September 9, 1924. Assigned to General Electric Company, of New York.

ELECTRON DISCHARGE DEVICE, in which the elements are arranged within an evacuated vessel in such manner that the controlling electrode is located about the cathode and within electron receiving relation therewith. The anode is enclosed by the cathode. The anode may consist of a rod extension within the cathode while the control electrode may comprise a small cup enclosing the other electrodes.

1,508, 615—E. C. Rowley, filed December 27, 1922, issued September 16, 1924.

CRYSTAL DETECTOR FOR RADIO APPARATUS, in which a multipoint contact element is provided, consisting of a conducting disk, a clamping disk and securing means for locking the disks together, in order to hold a plurality of cat whiskers between the disks. The plurality of cat whiskers are placed in contact with the surface of a sensitive crystal element.

1,508,893—F. G. Mitchell, filed June 24, 1922, issued September 16, 1924.

CRYSTAL DETECTOR FOR RADIO RECEPTION, in which a pair of plates supported by a base is arranged with serrated edges between which a sensitive crystal is gripped, thereby providing a rectifying system.

1,509,139—D. Grimes, filed April 21, 1922, issued September 23, 1924. Assigned to American Telephone and Telegraph Company, New York.

RADIO RECEIVING APPARATUS, in which the filaments of the electron tube amplifier are heated from an alternating current of relatively low frequency. A coupling is provided between the filament circuit of the tubes and the source of alternating current, including a condenser offering small impedance to radio frequencies and large impedance to other frequencies and a shunt associated with the coupling and arranged to be anti-resonant to radio frequency, but resonant to lower frequency.

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1,510,341—J. A. Proctor, filed June 4, 1921, issued September 30, 1924. Assignor to Wireless Specialty Apparatus Company, Boston.

ELECTRICAL APPARATUS, forming a quick break-in system, for placing receiving apparatus at a radio station in condition of operation intermediate the transmitting periods. The apparatus consists of a switch, the contacts of which are operated within a vacuum. The contacts are positively opened during the transmitting intervals disconnecting the receiver from the antenna.

1,510,624—A. J. Kloneck, filed March 31, 1920, issued October 7, 1924.

ELECTRICAL POWER TRANSMISSION BY RADIATION, for the purposes of driving machinery at distant points from one or more central power stations. The patent describes a system for transmitting and receiving light and motor power, comprising a plurality of independently operating generating dynamos arranged at remote points from one another and adapted for cooperatively radiating electrical energy of one character and of a frequency below 200 cycles. Independent radiating circuits are provided for the separate dynamos and each of the radiating circuits are tuned to a corresponding frequency.

1,510,741—E. G. Danielson, filed October 30, 1919, issued October 7, 1924. Assigned to Gray and Danielson Manufacturing Company, San Francisco.

ROTARY SPARK GAP APPARATUS for a radio transmitter, wherein the rotor is driven by a rotable shaft passing thru a rotary electrode drum. The stationary supporting drum is provided with insulators projecting therein, which support stationary electrodes with pairs thereof, interconnected by arc-shaped members upon which the stationary electrodes may be relatively adjusted with respect to the rotary electrodes.

1,510,792—E. Merritt, filed July 1, 1921, issued October 7, 1924.

METHOD OF AND MEANS FOR DETERMINING PHASE DIFFERENCE, whereby the direction of the waves may be determined. The fundamental principle involved is the determination of phase differences between trains of waves or vibrations, by means of the phase difference between the beats that are produced when these wave trains or vibrations are combined, by the process of heterodyning with another wave train or vibrations of a different frequency.

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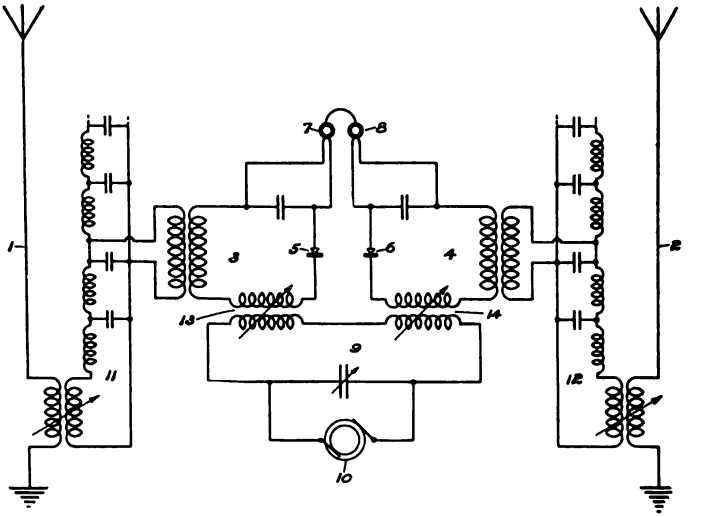
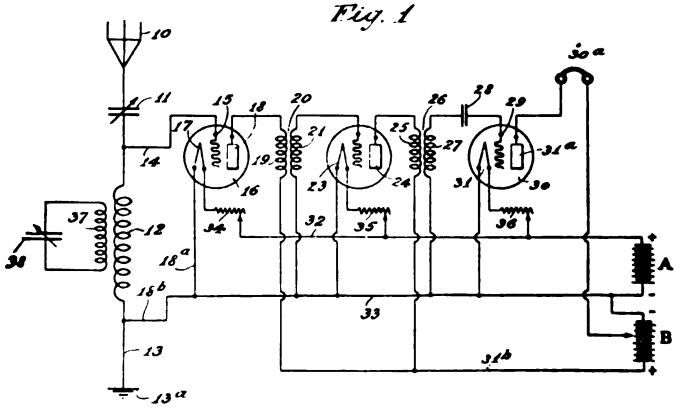


NUMBER 1,510,792—Method of and Means for Determining Phase Difference

1,510,799—J. H. Rogers, filed July 12, 1923, issued October 7, 1924.

LOOP AERIAL, in which all of the turns of the loop lie substantially in the surface of a polygonal tore. The loop is constructed on a frame having arms at each end, on which a circular disk-like member is provided. The circular members are apertured to receive the turns of the loop in such position that all of the turns are disposed symmetrically with respect to the others.

1,510,945—E. B. Lewis, filed August 8, 1922, issued October 7, 1924.



NUMBER 1,510,945—Radio Receiving System

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RADIO RECEIVING SYSTEM, in which undesired locally generated "feed back" currents occurring in the electron tube circuits are suppressed and thus prevented from building up to an injurious degree. An absorbing device for the undesired oscillations fed back into the grid circuit from the associated tube circuits is provided. The device comprises an inductance coil in the grid circuit, including an outer stationary ring in series therewith. The stationary outer ring has an inner-closed ring mounted on an axis in the plane of the outer ring and adapted to be rotated on its axis, whereby the energy of said oscillations will be selectively absorbed by induction and dissipated as heat.

Re. 15,924—E. A. Sperry, filed March 12, 1920, issued October 7, 1924. Assignor to Sperry Gyroscope Company, Brooklyn, New York.

RADIO REPEATER SYSTEM for automatically transmitting the readings of an indicating member by radio to a distant point. The invention is described as applied in automatically transmitting by radio to an aircraft or other distant point the readings of a compass, fire control instrument, or other indicator on board a ship, whereby the ship's heading or the bearing of a target observed from the ship may be constantly communicated to such distant point. Such a system has especial uses in the aerial ordnance control for heavy guns where the target is only visible from airplanes, which signal to the ships or other firing station the bearings of the target. The ship's compass is used to control the radio transmitter and at the receiver an indicator is actuated to reproduce the compass reading.

1,511,935—E. A. Bayles, et al, filed July 20, 1921, issued October 14, 1924.

ELECTRICAL CONDENSER, in which a liquid impregnating dielectric is used and the condenser units are formed of laminae of paper and metal. The dielectric is wound around a layer of corrugated material which spaces the condenser sections apart. By reason of the separation of the condenser units by the corrugated material a high degree of insulation against breakdown is obtained.

1,512,398—R. C. Browne, filed October 14, 1922, issued October 21, 1924.

AIR CONDENSER, having a plurality of movable vanes and a locking member associated with one of the vanes adapted to co-

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operate with a desired number of the other movable vanes to control their operation in unison. The condenser may be varied over small ranges of capacity by moving one of the vanes independently of the others and then moving all of the vanes or a portion thereof simultaneously.

1,512,824—H. M. De Bellescize, filed August 29, 1921, issued October 21, 1924.

RADIO TELEGRAPH AND TELEPHONE RECEIVER, which is substantially protected against static and strays by a differential circuit, including two opposing resonators. The incoming signaling energy is impressed upon a detector and then upon a pair of circuits, tuned to slightly different tone frequencies. The circuits are connected in opposition and deliver energy to a telephone receiver circuit, connected with both of the opposition circuits in such manner that persistent signals may actuate the responsive device while strays may be canceled.

1,512,960—J. Weinberger, filed October 11, 1921, issued October 28, 1924.

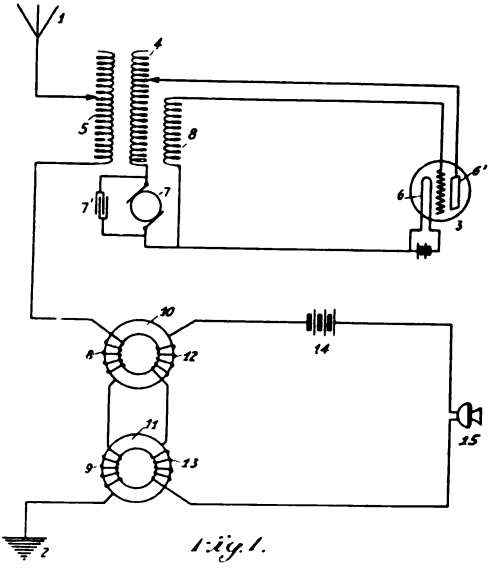

Fig. 1.

NUMBER 1,512,960—Radio Signaling System

RADIO SIGNALING SYSTEM, employing a modulation circuit for controlling the amplitude of radio-frequency signaling energy by means of an arrangement in which the variation of an audio-fre-

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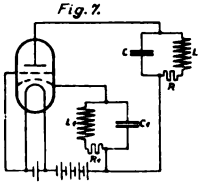
quency magnetization circuit having an iron core, co-operates with a radio-frequency winding on the core to greatly vary the effective resistance of the latter. The change in the resistance of the radio frequency winding is due to the change of the iron losses of the core when the direct current magnetization is varied.

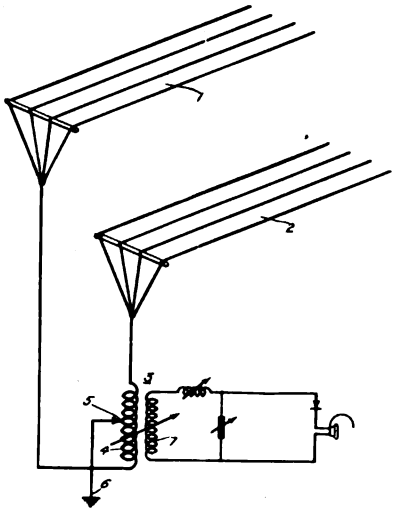
1,513,010—H. Rukop, filed May 3, 1922, issued October 28, 1924.



NUMBER 1,513,010—Arrangement for Producing Electrical Oscillations

ARRANGEMENT FOR PRODUCING ELECTRICAL OSCILLATIONS, wherein an electron tube, provided with a special positive electrode is employed, connected in an oscillatory circuit. The tube contains a cathode, a space-charging positive electrode, a grid and an anode arranged in the order named. A circuit connects the space-charging electrode and the anode with the cathode. A grid circuit and an anode circuit is provided, the positive space-charging electrode operating to influence the negative characteristic of the tube to maintain the circuits in oscillation.

1,513,223—F. Conrad, filed October 17, 1918, issued October 28, 1924.



NUMBER 1,513,223—Receiving Circuit for the Elimination of Static Disturbances

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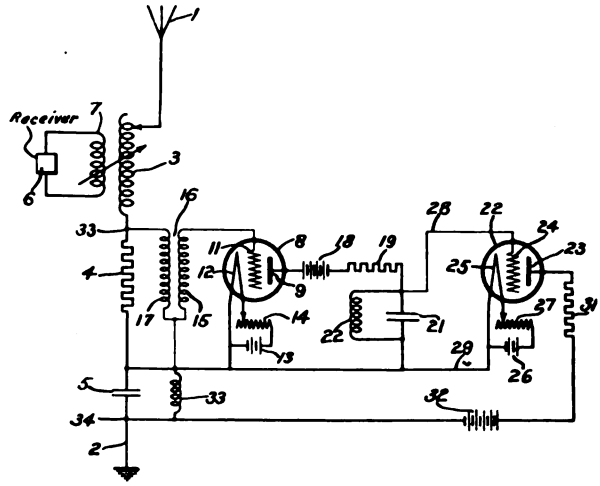
RECEIVING CIRCUIT FOR THE ELIMINATION OF STATIC DISTURBANCES, employing two antennas associated with the receiving apparatus in such manner that substantially equal charges imparted thereto by static influences cancel each other within the receiving apparatus, thus being rendered imperceptible to the operator. The antennas are supported at different heights so that the signal impulses will be stronger in one antenna than in the other. The unequal amounts of energy imparted to the antenna by incoming signal impulses fail to cancel each other and thus a residual effect is imparted in the receiving apparatus which may be amplified if desirable, in order to render it more clearly perceptible to the operator.

1,513,286—J. Slepian, filed November 16, 1921, issued October 28, 1924.



NUMBER 1,513,286—Static Elimination System

STATIC ELIMINATION SYSTEM for radio receiving apparatus, wherein a resistor having such a value as effectively to damp the receiving system is connected in the antenna circuit. A non-resonant circuit is associated with the resistor, in such manner that the voltage drop therein may be substantially neutralized for current of predetermined frequency only whereby the system is screened against static.

1,513,324—S. M. Kintner, filed October 25, 1921, issued October 28, 1924. Assignor to Westinghouse Electric and Manufacturing Company.

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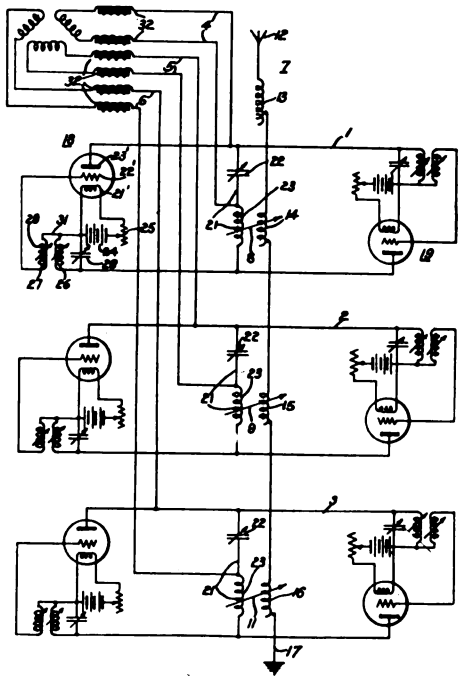

NUMBER 1,513,324—Polyphase Plate Circuit Excitation System

POLYPHASE PLATE CIRCUIT EXCITATION SYSTEM for electron tubes connected to form an oscillation generator. A circuit arrangement is shown for the connection of a plurality of tubes, wherein the plates are energized from three phase alternating currents of commercial frequency. The invention describes a transmission system, wherein commercial three phase alternating current is placed directly upon the plates of the several tubes which supply oscillations to a common circuit associated with the radiation system.

1,513,326—W. C. Lamphier, filed February 9, 1923, issued October 28, 1924.

RADIO DETECTOR, consisting of two contact devices, one of which is a steel-wool sensitive contact member and the other graphite. The steel-wool is mounted in the usual detector cup and the graphite is adjusted in contact therewith to obtain the desired rectifying characteristic.

1,513,561—F. Schroter, filed October 5, 1920, issued October 28, 1924. Assignor to Gesellschaft für drahtlose Telegraphie m. b. H. Berlin, Germany.

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GAS RELAY, comprising an evacuated vessel containing an anode, a grid and a cathode, which consists of an alkaline metal together with an inert mixture of a gaseous medium in the tube.

1,513,707—J. H. Hammond, Jr., original filed August 5, 1919, issued October 28, 1924.

Fig. 1

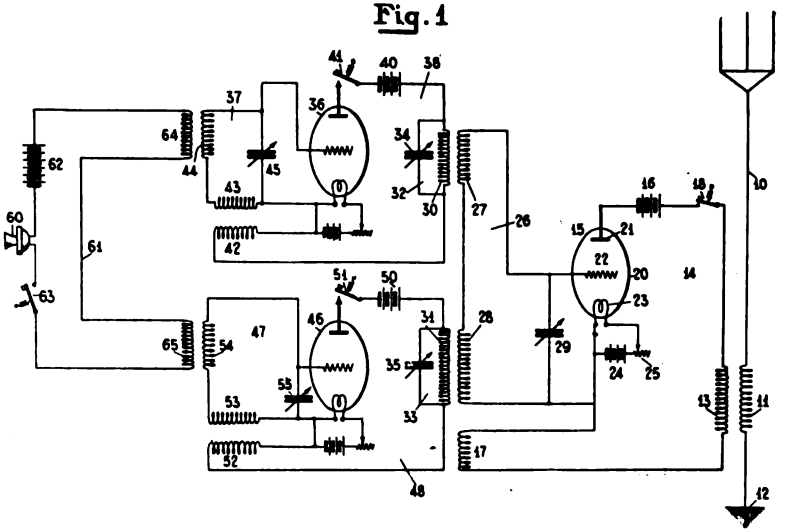
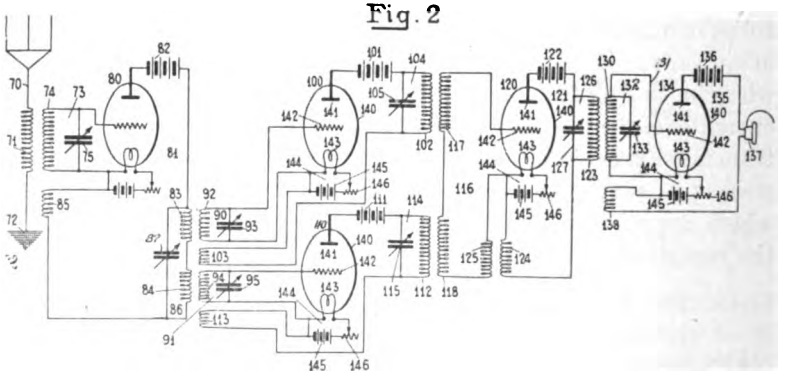


Fig. 2



NUMBER 1,513,707—Transmission and Receiving System

TRANSMISSION AND RECEIVING SYSTEM for secret communication by radio, wherein the transmitted signals consist of a series of impulses having impressed thereon a plurality of series of periodic modifications of different frequencies which are received and combined at the receiving station to produce an intelligible signal.

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VOLUME 12 FEBRUARY, 1924 NUMBER 1

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PROCEEDINGS
of
The Institute of Radio Engineers

EDITED BY
ALFRED N. GOLDSMITH, Ph.D.

PUBLISHED EVERY TWO MONTHS BY
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Acceptance for mailing at special rate of postage provided for in Section 1103, Act of October 3, 1917, authorized October 7, 1918.

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C-12—Similar to C-11 with standard base.
$5.00

C-11—1.1 Volt .25 amp. Dry Battery Det. and Amp. Special base.
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C-299—3 Volts .06 amp. Dry Battery Det. and Amp.
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For Every Type of Receiving Set

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C-12 is identical to C-11, but is mounted on a base for use in sets equipped with standard sockets without the aid of special adaptors.

Whenever storage battery supply is available, the C-300 is the most sensitive detector for reception of distant and weak signals.

Under the same condition C-301 A is the best tube for amplification at either radio or audio-frequency.

PATENT NOTICE

Cunningham tubes are covered by patents dated 2-18-08, and others issued and pending. Licensed for amateur, experimental and entertainment use in radio communication. Any other use will be an infringement.

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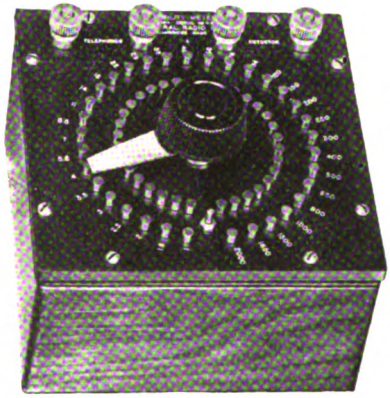
AUDIBILITY METER



Type 164

AUDIBILITY METER

Price, $36

The ear alone cannot be depended upon to estimate the intensity of received signal strength. Some auxiliary standard of reference must be used.

One of the more common methods is to shunt the telephone receivers. As the receivers are shunted, however, a series resistance of such a value should be added to the circuit that the impedance will remain unchanged.

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Send for Bulletin R

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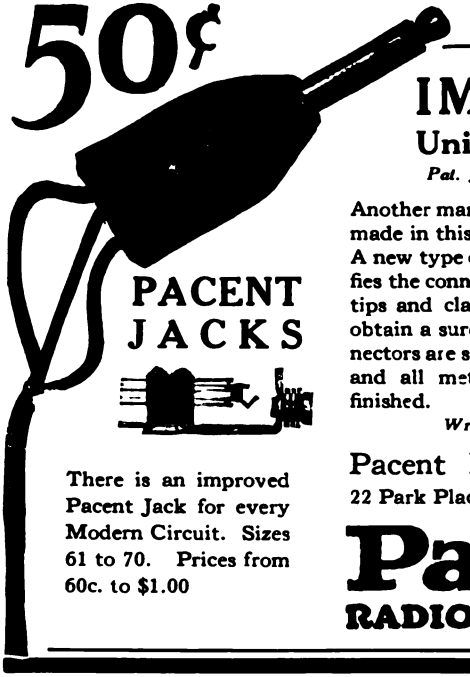

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Pat. Jan. 8, 1924, Other Pat. Pend.

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U.S.A.

Dear Mr. James,

Back home since four days only I have reinstalled my station for 100 meter work.

I am glad to say the GREBE CR-13 which I brought back has already done splendid work. I tuned down to 100 meters for the first time this morning at 0010 G.M.T. and immediately heard KDKA. I followed him until 0.00 and all the time his carrier wave was good, he could have been easily read in telegraphy. At times excellent orchestral music and some words were plainly heard in spite of heavy static and arc interference.

The distance from Pittsburgh to Nice must be about 7,000 kilometers (roughly 4,400 miles); I have all reasons to think this is the world record distance transmission on 100 meter wave.

Please express my hearty congratulations to the GREBE Co. for their wonderful receiver.

Very sincerely yours.

Léon Deloy
8AB

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Doctor Mu

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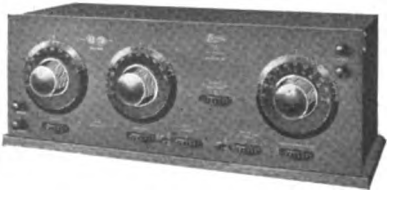

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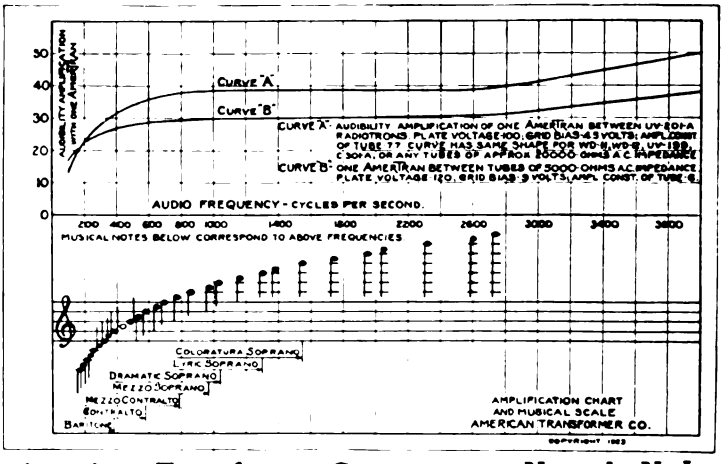
AUDIBILITY AMPLIFICATION OVER ONE AMERTRAN
50
40
30
20
10
0
CURVE 'A'
CURVE 'B'
CURVE 'A' AUDIBILITY AMPLIFICATION OF ONE AMERTRAN BETWEEN UV-201-A RADIOTRONS. PLATE VOLTAGE 100, GRID BIAS -4.5 VOLTS; AMPL. CONST. OF TUBE 7.7. CURVE HAS SAME SHAPE FOR WD-11, WD-12, UV-199, C 301-A, OR ANY TUBES OF APPROX 30,000 OHMS A.C. IMPEDANCE
CURVE 'B' ONE AMERTRAN BETWEEN TUBES OF 30,000 OHMS AC IMPEDANCE PLATE VOLTAGE 120, GRID BIAS -9 VOLTS, AMPL. CONST. OF TUBE 6
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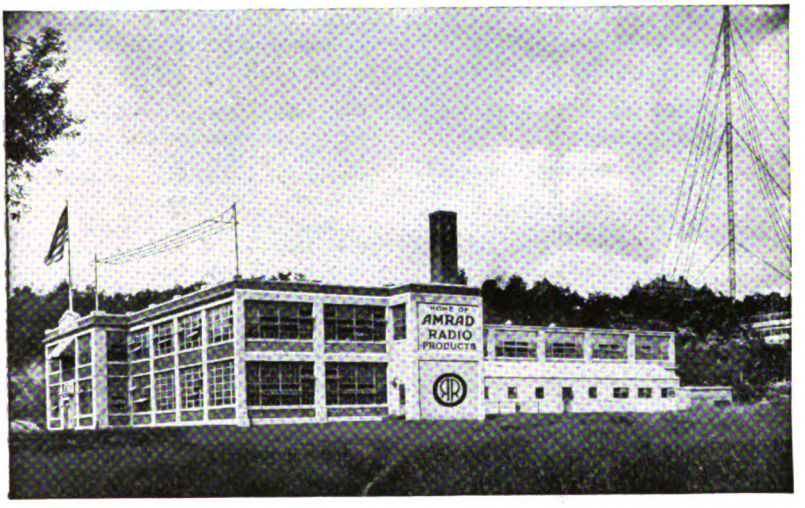

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A Radio Statement to the Public

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Briefly stated, there is today a necessity for

—*A radio receiver providing super-selectivity* —the ability to select the station you want—whether or not local stations operate. A selectivity which goes to the theoretical limits of the science.

Super-sensitiveness — meaning volume from distant stations—along with selectivity

—*Improved acoustics*—more faithful reproduction of broadcasted voice and music than has ever been possible before.

—*"Non-radiating" receivers*—a new development, a type of receiver which, no matter how handled, will not interfere with your neighbor's enjoyment.

—*More simplified operation*—a super-receiver requiring no technical skill, thus making the greatest achievements of entertainment immediately available to all members of the family.

—*A receiver for the apartment house* and populated districts, requiring neither aerial nor ground connection.

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Third—that the regenerative receiver could be markedly improved—providing selectivity, sensitiveness and simplicity of operation hitherto deemed impossible.

Fourth—that the Super-Heterodyne—the hitherto complicated device requiring engineering skill to operate could be vastly improved—improved in sensitiveness and selectivity—and simplified so that the very novice and the layman could enter new regions of entertainment and delight.

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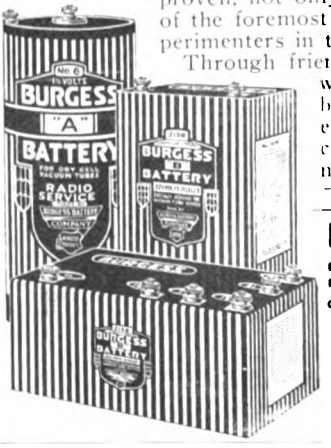

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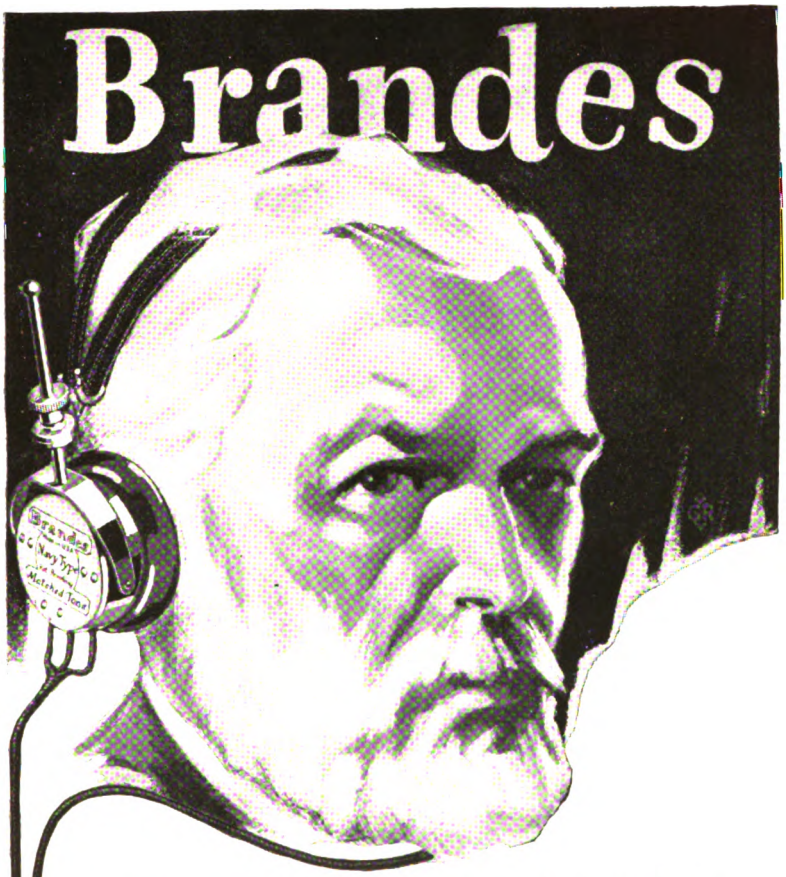

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As you know, the MacMillan Expedition station WNP is at present frozen in somewhere near the North Pole. We're all trying to pick up their nightly messages. Once in a while we succeed. But to the little Ace Type V and Ace Type 3B, operated by Mr. Weeks, 9 DKB, goes the distinction of being the only consistent relay point of WNP.

Here's what Mr. Weeks says:

Nov. 21, 1923.

"The little Ace Type 3B isn't so bad for selectivity, cutting out a ten watter, radiating 3 amps, and bringing in the louder B. C. Stations. Hung up a little record with one last night. Heard WNP and worked him, using 50 watts and the 3 B. Hi."

Dec. 11, 1923.

"Worked WNP the other morning and took 1,500 words NANA for him. Took a message from him the morning before that was destined for Sommerville."

This is the first NANA message from WNP sent direct to the U. S. It took an hour, and the Ace Type 3B, on which it was received, brought it in QSA without a miss.

Dec. 19, 1923.

"Took twelve messages from MacMillan two hours, using Ace Type Five."

Now do you believe us? If not we can give you plenty more evidence.

But maybe this isn't news to you. Perhaps you have had success in bringing in WNP. If not, keep on trying. Your Ace Model V or Ace 3B might not bring them in the first time you try, but if they can do it in Minot, N. D., they can probably do it for you.

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For LEAD IN, No. 17 Rubber Insulated Copperweld Wire
For GROUND Wire, same as Lead-In Wire
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MAGNAVOX instruments are never subject to those interferences which, at critical moments, are so apt to mar the performance of ordinary radio reproducers.
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New York—San Francisco

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CANADIAN NATIONAL CARBON CO., Limited.
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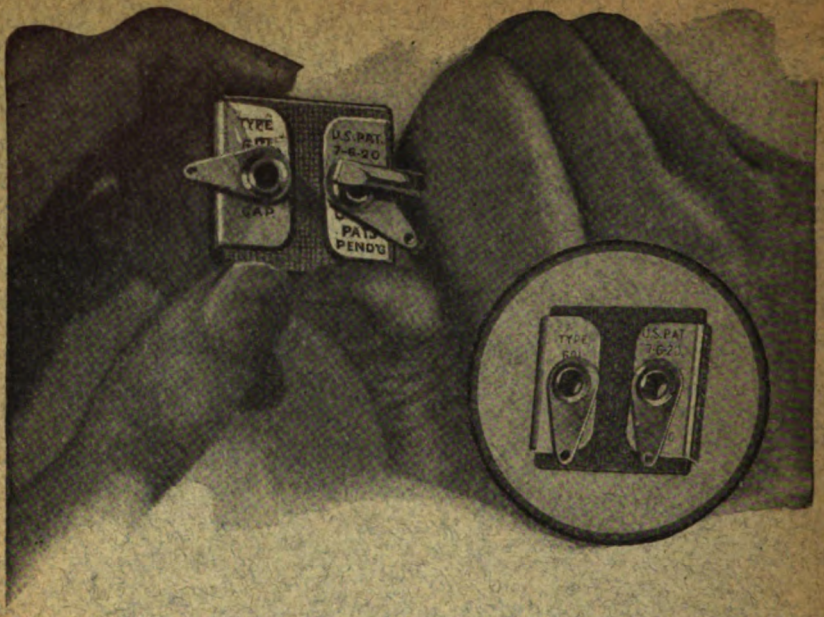

If the soldering tabs are not needed, simply swing them out of the way, as here shown

Micadon Type 601 — *now has swinging soldering tabs*

The new swinging soldering tabs with which the Dubilier Micadon is now provided, facilitate mounting.

The hot soldering iron need not touch the Micadon directly. Hence the capacity is not affected by heat.

If Dubilier Micadons are to be mounted with screws, simply swing the soldering tabs out of the way and use the eyelets.

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ALFRED N. GOLDSMITH, Ph.D.

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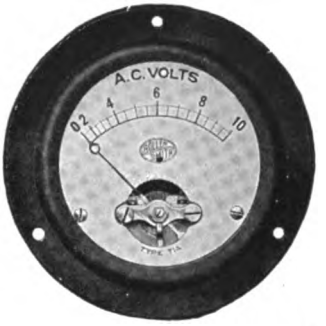

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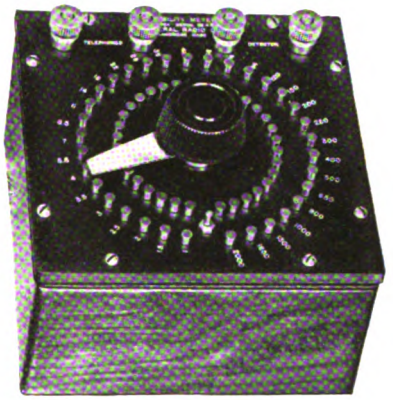

Type 164

AUDIBILITY METER

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The ear alone cannot be depended upon to estimate the intensity of received signal strength. Some auxiliary standard of reference must be used.

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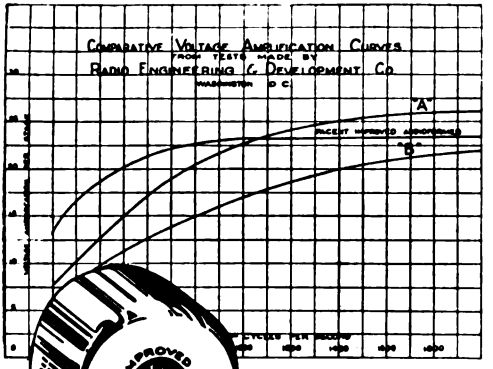
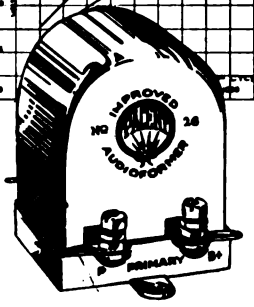


Comparative voltage amplification curves from tests made by the Radio Engineering and Development Company of Washington, D. C.



The Chart tells the story

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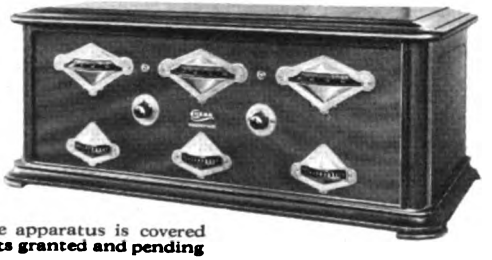

All Grebe apparatus is covered by patents granted and pending

THE high degree of selectivity and over-all efficiency attained in the design of the Grebe Synchrophase is rivalled only by its rare craftsmanship and thorough ease of dependable operation.

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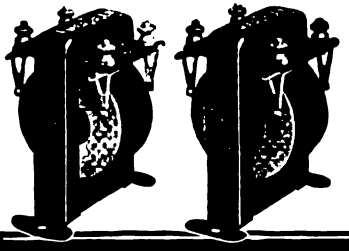

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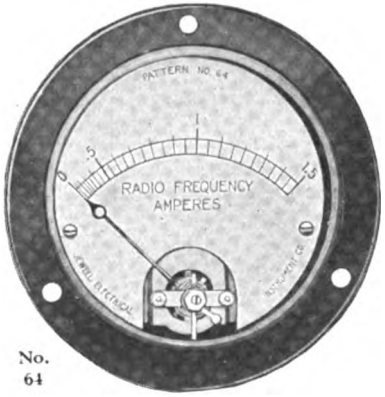

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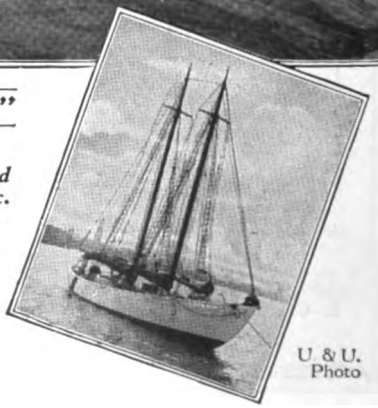

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It needs no extra batteries

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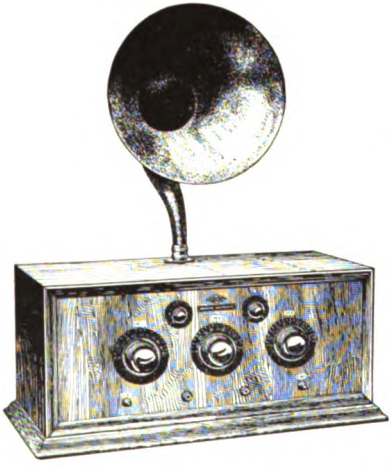

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There can be no gift with greater possibilities for continued happiness than a Crosley set. It carries Christmas along through the year, continually giving new thrills and happiness, and bringing pleasant thoughts of the giver.

It is a delight to operate a Crosley. The immediate response to the turn of the dials; the clearness of reception from far distant points; the real ease with which local stations may be tuned out; all help to make Crosley reception distinctive and exceptionally pleasurable. The very low cost at which this really remarkable radio performance can be obtained places Crosley sets within the reach of all—the ideal Christmas gift.

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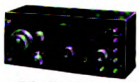

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That's what *one* expert says of De Forest Tubes. Many other radio engineers have appreciated and endorsed their uniformity of excellence. The De Forest Tubes are up to the standard their name demands. They stand the severity of experimental work and laboratory tests. They deserve the leadership Lee De Forest first gave them in 1906.

Agreement is general about this. And the De Forest Company guarantees its products and backs this guarantee with a liberal replacement policy. No tube can get out of the De Forest factory without going through a whole series of tests and inspections. But the final and most significant test of all is performance when used by the purchaser.

There are two types, and only two are necessary. They are shown below — the DV-3 for dry batteries and the DV-2 for storage. The DV-3 has an average mutual conductance of 460 micro-mhos; average amplification constant (or Mu) 6; plate impedance 13,000 ohms; a good radio and audio frequency amplifier and a good detector in standard regenerative circuits. It is remarkably non-microphonic. The DV-2 for storage batteries is made particularly for power amplifier work and is developed for all usual circuits. Conductance 720 micro-mhos; amplification constant (or Mu) 7.2; plate impedance 10,000 ohms.

These tubes are sold by authorized dealers only and are made by the makers of De Forest D-12 Radiophones and all radio parts.

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Jersey City, N. J.

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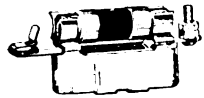

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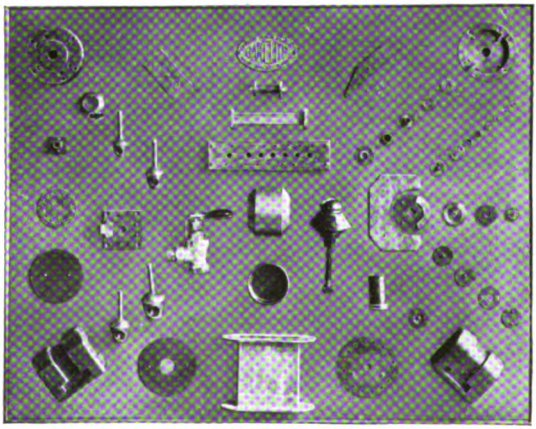
BOONTON BAKELITE first opened the way to low-loss parts at low prices.

BOONTON BAKELITE
has always meant
GENUINE BAKELITE



The first Bakelite parts *molded at Boonton in 1909*

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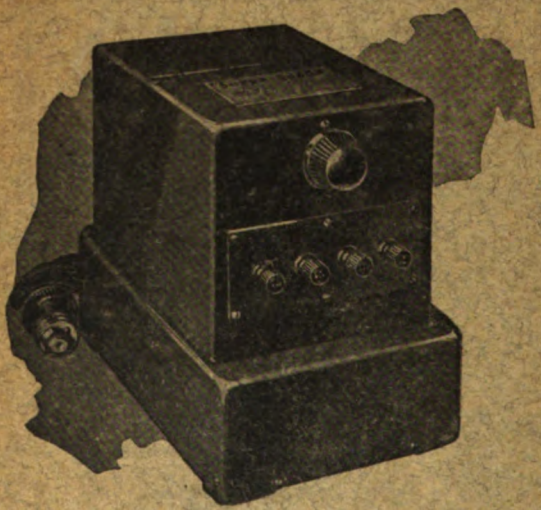

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